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THE JOURNAL OF RADIO RESEARCH & PROGRESS

JANUARY 1950

VOL. XXVII TWO SHILLINGS AND SIXPENCE No. 316

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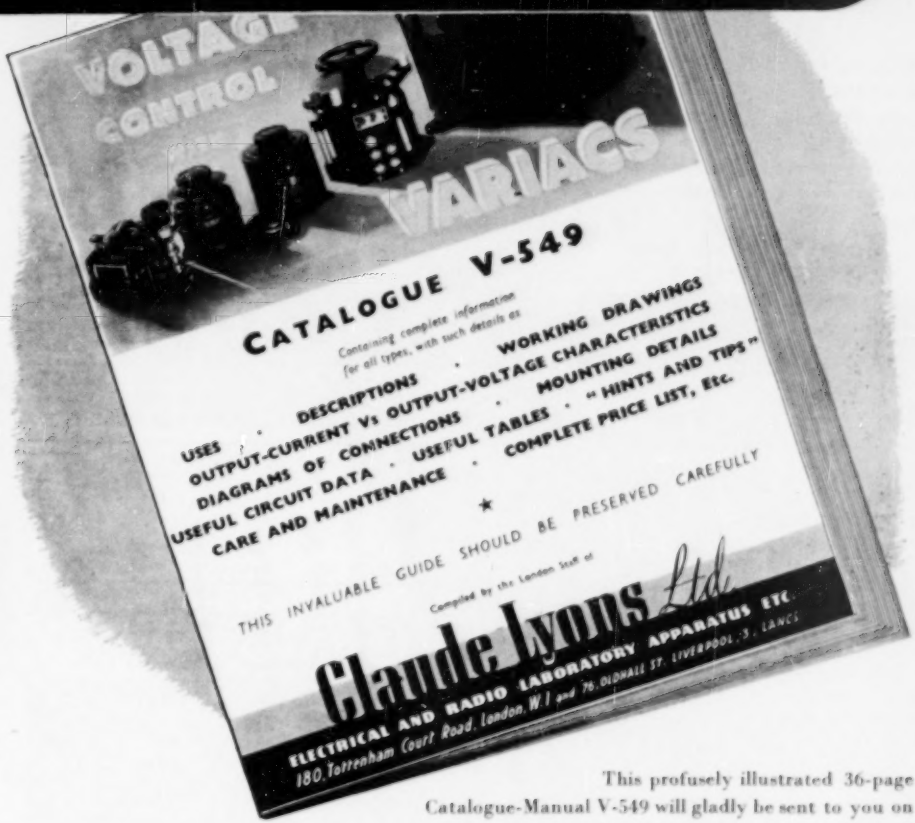
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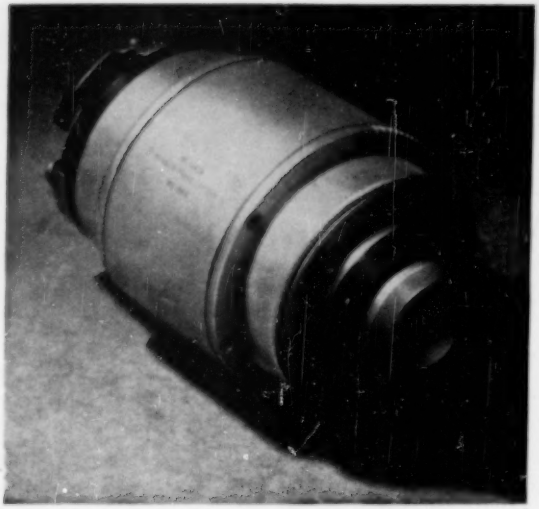

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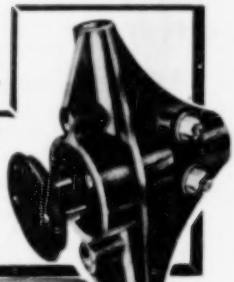

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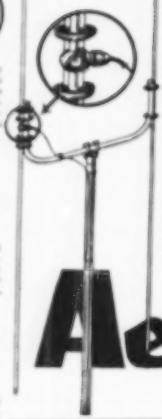

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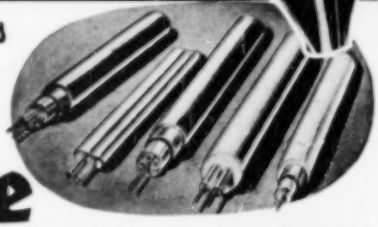

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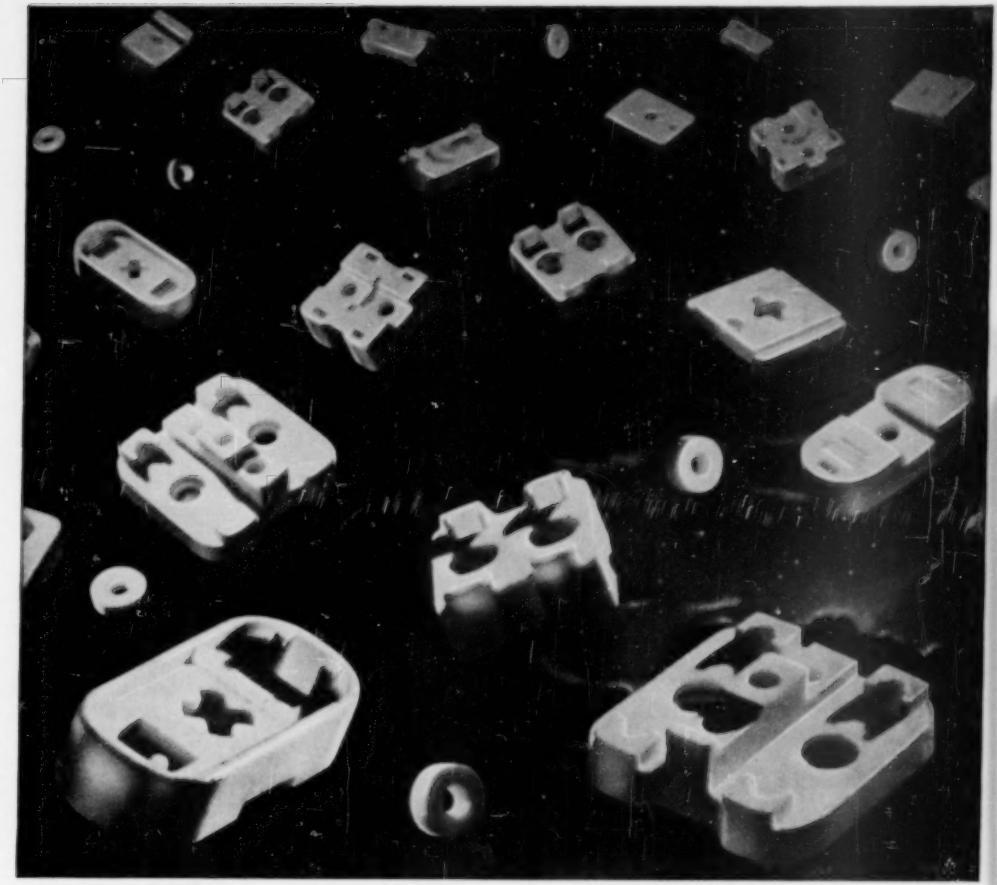

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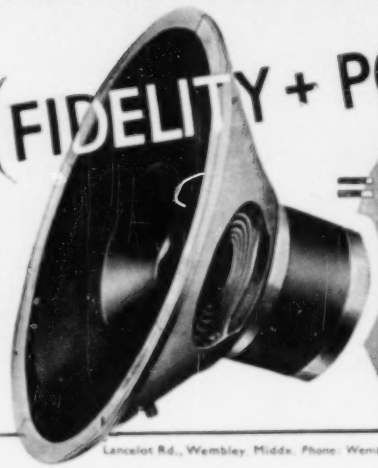

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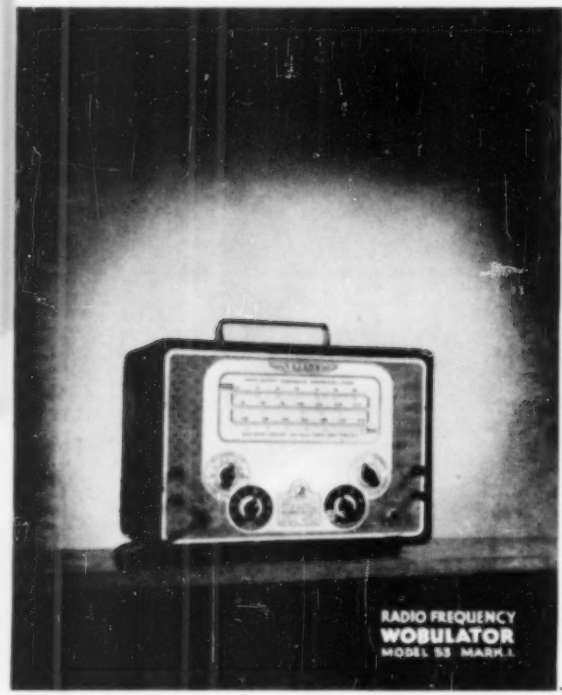

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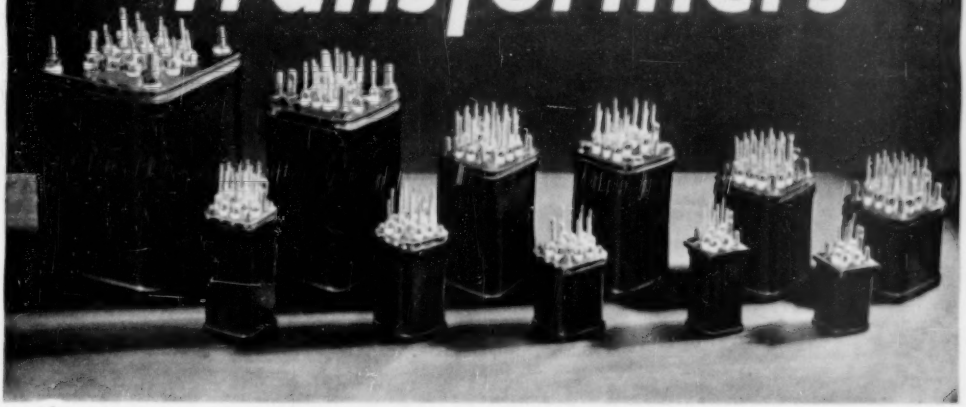

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MODEL · RM 175-L

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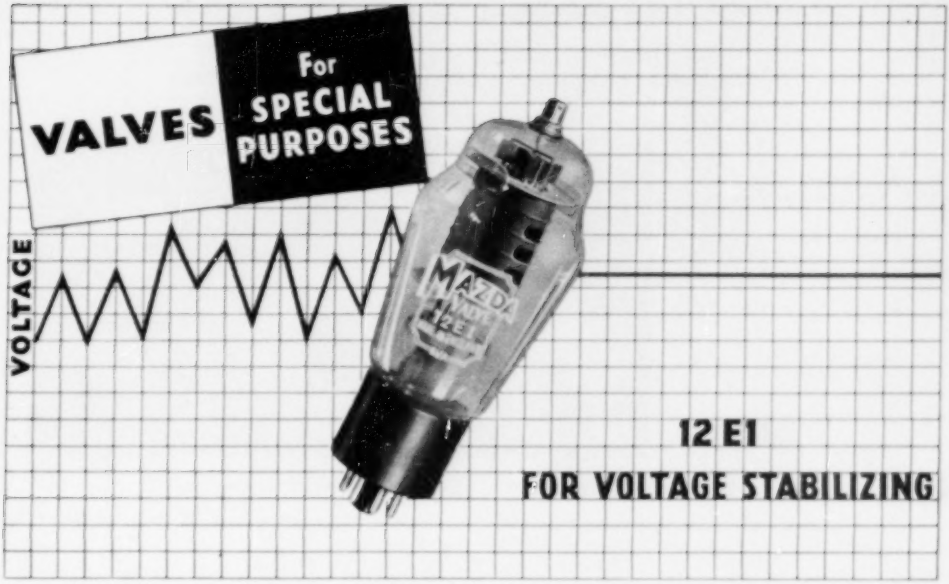

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Max. Grid 1 to screen volts	... ...	400
Max. Anode dissipation (watts)	... ...	35
Max. Screen dissipation (watts)	... ...	5.0
Max. Cathode current (mA)	... ...	300
Max. Heater to Cathode volts	(Cathode positive)	300

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Maximum overall length (mm.)	... ...	150
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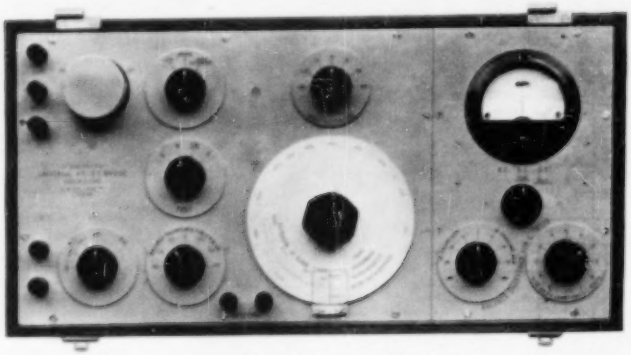

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JANUARY 1950

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- Good frequency multiplier.
- Sections may be used in cascade.

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I_f	0.68 A
f	186 Mc/s
V_a	400 V
I_a	2x40 mA
*Pload	16.5 W

* For circuit efficiency = 85%

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WIRELESS ENGINEER

Vol. XXVII JANUARY 1950 No. 316

The Rationalized M.K.S. System

THIS was the title of the Editorial of November 1948. In it we made a statement which might cause some misunderstanding. We said "It should be noted that rationalization does not affect B, D or $\mathcal{E}$, but only H, μ and κ." We have been asked how κ could be affected if D and $\mathcal{E}$ are not—a very reasonable question. The difficulty arises over the definition of D; Maxwell, who introduced the conception of displacement, was quite definite about it, and stated that "the displacement through a given surface is the quantity which passes through it." The quantity passing through unit area is sometimes called the displacement density and sometimes simply the displacement. In the British Standards Glossary of Terms used in Electrical Engineering electric displacement is given as an alternative name for electric-flux density, which is defined as the quantity of electricity displaced across a given area in a dielectric.

In the m.k.s. system of units the unit of displacement density, is the coulomb per square metre, and the unit of electric force $\mathcal{E}$ is the volt per metre. Now neither of these would be affected by rationalization, for the simple reason that D was introduced by Maxwell in the rationalized form. If we write $D = \kappa_0\mathcal{E}$ for empty space then $\kappa_0 = 1/(36\pi \times 10^9)$ which is its rationalized value, but if for any reason one prefers to write $D = \kappa_0\mathcal{E}/(4\pi)$, then κ_0 will be increased to $1/(9 \times 10^9)$ which is its unrationalized value. In the c.g.s. electrostatic system the value of $\mathcal{E}$ at a distance r from a charge q was $q/(\kappa r^2)$ and the value of D was $q/(4\pi r^2)$, so that $D = \kappa\mathcal{E}/(4\pi)$. κ is here the unrationalized c.g.s. value; that is, unity. It can be rationalized by writing $D = \kappa_0\mathcal{E}$ where $\kappa_0 = \kappa/(4\pi)$.

The possible misunderstanding to which we referred is due to the fact that several authors of textbooks have modified Maxwell's definition of displacement and have de-rationalized it in order to bring it into line with the magnetic formula $B = \mu H$. Instead of following Maxwell and making a unit tube of displacement end on unit charge, they take as a unit tube, one ending on a charge of $1/(4\pi)$ of a unit. Those who have done this will, of course, have to retrace their steps when introducing rationalization.

Those who have de-rationalized Maxwell's displacement in order to assimilate the electric and magnetic expressions, obtain for the electric energy per unit volume the formula $D^2/(8\kappa_0\pi)$, where $\kappa_0 = 1/(9 \times 10^9)$, giving exact parallelism with the formula $B^2/(8\mu_0\pi)$ for the magnetic energy per unit volume. Using Maxwell's rationalized displacement the formula is $2\pi D^2/\kappa_0$ which will be very familiar to most of our readers. If we now complete the rationalization by rationalizing κ_0 the formula becomes $D^2/(2\kappa_0)$, where $\kappa_0 = 1/(36\pi \times 10^9)$. D is here in coulombs per square metre and the formula gives the energy in joules per cubic metre. Similarly, in the rationalized system the magnetic energy in joules per cubic metre is equal to $BH/2$, which for empty space becomes $B^2/(2\mu_0)$, where B is in webers per square metre, H in ampere-turns per metre, and $\mu_0 = 4\pi/(10^7)$.

G. W. O. H.

ELECTRICAL NOISE

Experimental Correlation between Aural and Objective Parameters

By D. Maurice, Ing.E.S.E. (Paris), A.M.I.E.E., G. F. Newell and J. G. Spencer

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1. Introduction

MANY types of electrical machinery cause radio interference because they make or break an electric current. The transient oscillation thus set up in neighbouring circuits may have Fourier components at audio and radio frequencies and these may be radiated or mains borne to audio- or radio-frequency receiving apparatus thus setting up a disturbance which, if the interference spectrum is uniform over the band of frequencies accepted by the receiver, is characterized mainly by the transmission properties of the receiver.

In order to suppress such interference it is necessary to decide subjectively how much of it is tolerable. The next step is to devise a measuring instrument capable of informing the engineer engaged in the suppression of interference when he has accomplished his task. This measuring instrument should thus have such characteristics as will enable its assessment of interference to simulate that of the person for whom the transmission is arranged. In the case of sound broadcasting it is the listener's ear which the interference-measuring device must simulate. The ideal interference-measuring set would thus consist of a typical receiver and loudspeaker and a human operator whose ear was tireless and never changing.

The tests to be described were undertaken with a view to ascertaining the law of variation of annoyance as the objective parameters of the interference were varied, and then the design or selection of a suitable measuring instrument.

In what follows the word loudness is invariably taken to mean the 'amount' of interference in the absence of broadcast programme. The word annoyance, on the other hand, is taken to mean the 'amount of disturbance' caused by the interference in the presence of a programme.

The impulsive type of interference, with which we shall usually deal, manifests itself in the receiver output in the form of successions of damped wavetrains. The time interval between them is usually much greater than their duration because the receiver bandwidth is much greater than the frequency of interruption of the current in normal types of electrical machinery such as motor-car ignition systems, electric motors, electric razors, thermostats, etc. It is thus convenient to idealize the interference into repeated Heaviside unit impulses. Such interference has only two parameters: the time integral, U, of its waveform and its pulse-recurrence frequency p.r.f.

The output of impulsive noise from the receiver will be a function not only of U and p.r.f., but also of the receiver modulation-frequency characteristic and any non-linear elements the receiver may contain. The shape of the receiver modulation-frequency characteristic has to conform to more stringent requirements than those which might be determined from impulsive-interference considerations and may be taken as more or less uniform over its acceptance band. We may now summarize and say that the significant parameters of the output noise are first, the peak or maximum value proportional to U and to receiver bandwidth; secondly, the duration of the noise phenomenon proportional to the reciprocal of receiver bandwidth, and thirdly the p.r.f. equal to that of the incoming interference. We have therefore to correlate annoyance with variation of each of these three parameters. The effect of the peak value on annoyance has been assumed a priori to be logarithmic so that we may start by measuring change of annoyance in decibels as units. This leaves bandwidth and pulse-recurrence frequency as suitable variables.

2. Experimental Arrangements

The experiments were based on a comparison between a standard noise and the noise whose parameters were varied. This method was chosen rather than working to an absolute value of annoyance such as adjusting the noise level until the annoyance conformed to some agreed standard such as 'tolerable' because it was thought to be quicker, to restrict the 'spread' or standard deviation of the listeners' results and to require less explanation to, and effort from, the listener himself. Fig. 1 shows a block schematic of the method adopted. Fig. 2 is a photograph of the 'listening room.' By means of a P.O. key the test listener could

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obtain any one of a number of noises, each adjustable in level by him, and compare them one at a time with the reference noise. The listener was allowed to adjust the level of reference noise at the start of the test provided that his adjustment brought it between the limits of 'just perceptible' and 'disturbing.' Measured on a Tannoy Phon Meter type 800 these limits were 70 phons less 30 db and 70 phons less 60 db. The noise level of the 'listening room' was about 70 phons less 34 db but this did not interfere with listening to noise levels lower than this, presumably due to spectrum differences between the test noises and the room background noise. The programme level was also set to the listeners wishes and was invariably between 70 and 80 phons.

3. The Noises

The programme used during annoyance tests was not specially selected but consisted of any one of the broadcast programmes existing at the time. It was conveyed to the listening room from the broadcast studio by land line and was thus in itself reasonably free from noise and amplitude or harmonic distortion.

The test noises were seventeen in number, excluding the reference noise. This consisted of random fluctuation noise in a band from 250 c/s to 10 kc/s. The reason for the bass cut below 250 c/s was that hum from source amplifiers

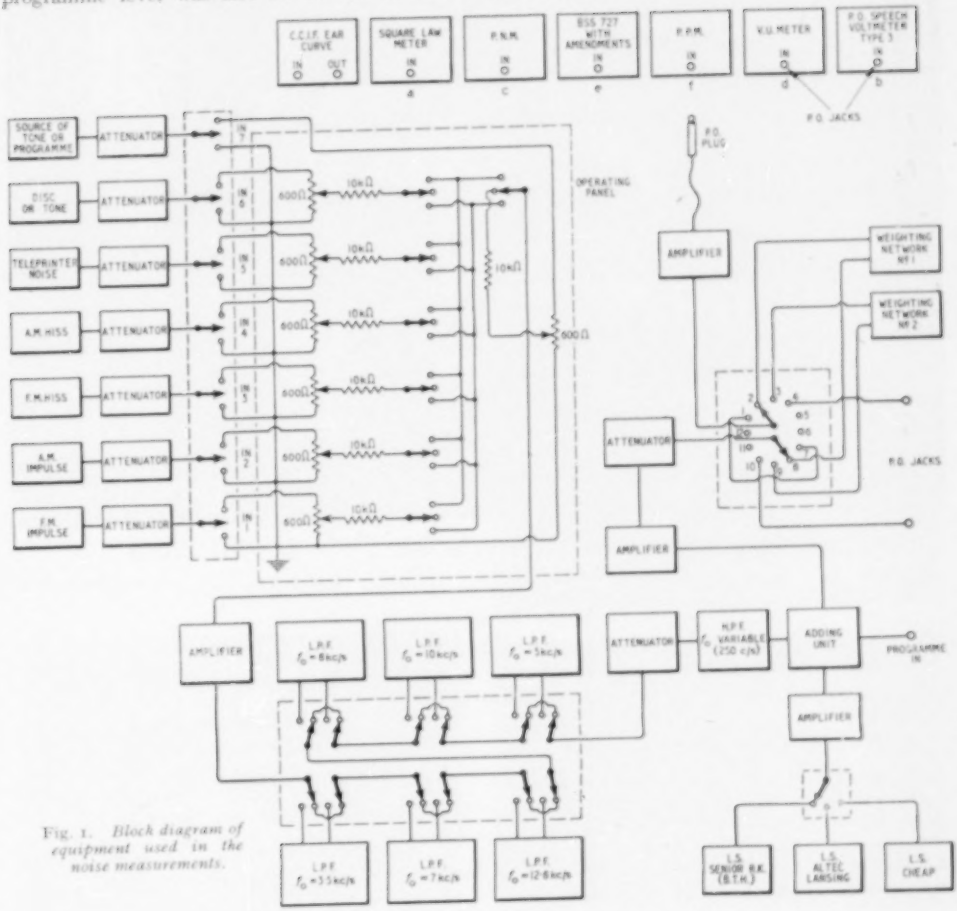


Fig. 1. *Block diagram of equipment used in the noise measurements.*

at the input of the a.f. chain caused trouble which was cured by the insertion of a 250-c/s high-pass filter. Fig. 3 (a) shews the spectrum. Noises one to eight were all restricted in spectrum to the band between 250 c/s and 10 kc/s.

tudes of successive output pulses have a random statistical distribution due to the random phase angles between each input pulse and a steady 'wanted' carrier wave.

Noise two was the output from a v.h.f. a.m.

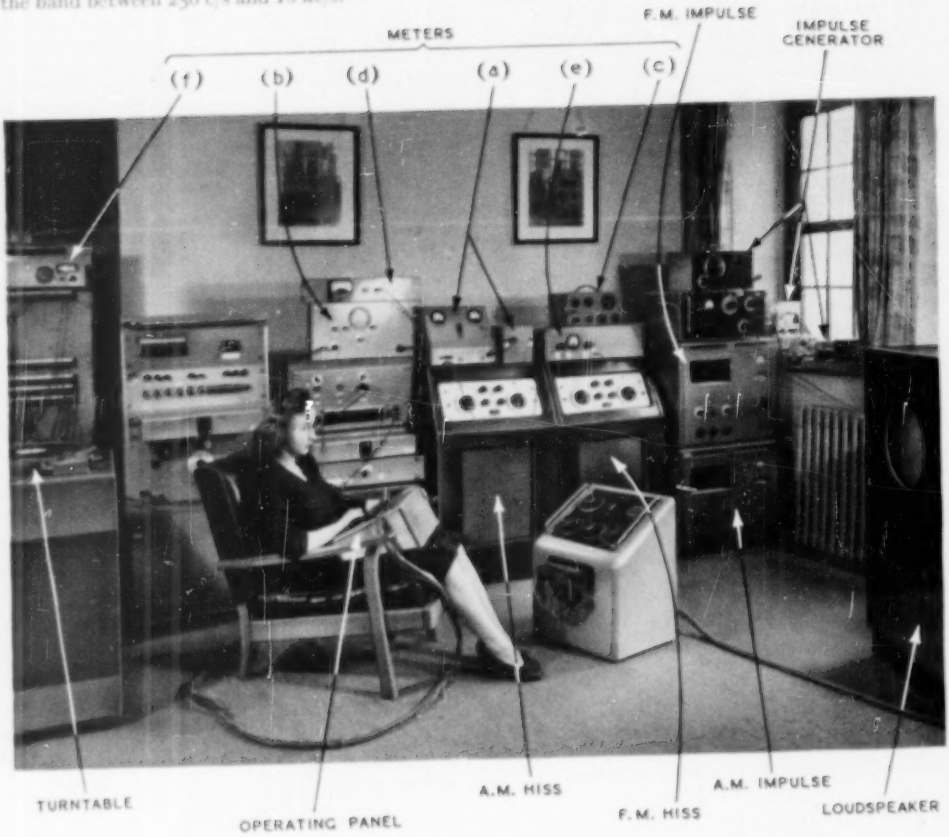


Fig. 2. *The listening room.*

Noise one consisted of the output from a v.h.f. f.m. receiver with 50 μs de-emphasis, the output being due to the application to its input of unit impulses repeated 100 times per second, Fig. 3(b). The receiver was operated just below the f.m./a.m. improvement threshold so that a small proportion of the output pulses had uniform rather than triangular spectra before de-emphasis, Fig. 3(c). Fig. 4 is a photograph of these output pulses, some being more or less unidirectional (uniform spectrum) and others bi-directional (triangular spectrum). The amplitudes of successive output pulses have a random statistical distribution due to the random phase angles between each input pulse and a steady 'wanted' carrier wave.

Noise two was the output from a v.h.f. a.m. receiver due to the same interference input as noise one. Its spectrum is shown in Fig. 3(a) and Fig. 5 is a photograph of it.

Noise three was random fluctuations emerging from a v.h.f. f.m. receiver with 50 μs de-emphasis. The spectrum is shown in Fig. 3(b).

Noise four was surface noise from a freshly-cut cellulose recording disc. The disc was cut in the absence of modulation on the cutter.

Noise five was teleprinter 'cross talk' obtained from a land line and recorded on a cellulose disc.

Noise six was 1000-c/s tone.

Noise seven was the same as noise one except that the p.r.f. was reduced from 100 c/s to 15 c/s.

Noise eight was the same as noise two but with a similar reduction of p.r.f.

Noises nine to seventeen were produced by feeding repeated unit impulses to a 30-c/s to 5-kc/s band-pass filter. The reference noise also had this bandwidth when these noises were used. The test arrangement was improved so that the hum was eliminated and the 250-c/s high-pass filter was no longer needed. The upper cut-off frequency of 5 kc/s was used in order to simulate conditions applying in the case of typical domestic broadcast receivers and also to be in close agreement with the bandwidth of a C.I.S.P.R. interference-measuring set, BS727. The spectrum of noises nine to seventeen is shown in Fig. 3(d). A photograph of noise nine is shown in Fig. 6. These noises differed only in p.r.f. Table I gives the p.r.f. of each noise.

4. Noise Meters

One way in which the law of variation of annoyance with variation of the objective noise parameters may be brought to light is to employ a number of different noise meters, each one having a known behaviour with respect to noise-parameter change. One might reasonably hope that one or more of the meters would closely follow the subjective results. This method was, in fact, adopted. Some of the meters were

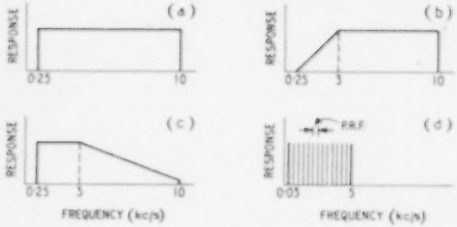


Fig. 3 (above). *Spectra of test noises; (a) reference noise and noise 2, (b) noises 1 and 3, (c) noise 1, (d) noises 9—17.*

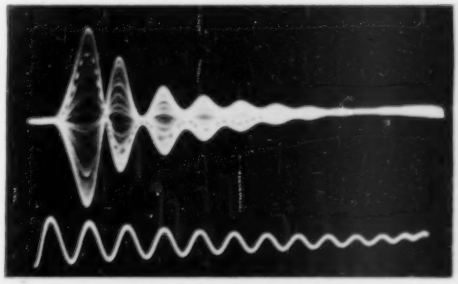
Fig. 4 (below). *Noise 1; dotted pulses are 'pops' due to operation of f.m. receiver just below improvement threshold. Untouched pulses are 'clicks' typical of f.m. reception. Amplitudes of successive pulses are random with time due to presence of a carrier. 10 kc/s timing wave shown below.*

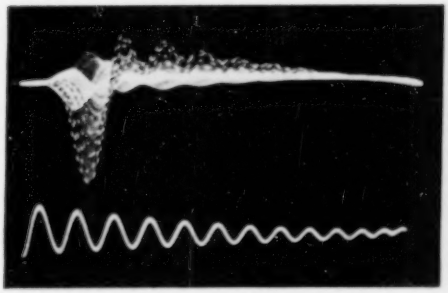
Fig. 5 (top right). *Noise 2; typical impulsive interference output from a.m. receiver. Amplitudes of successive pulses are random with time due to presence of a carrier. 10 kc/s timing wave shown below. One wavetrain dotted for clarification.*

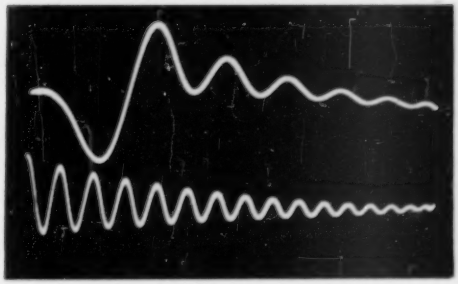
Fig. 6 (bottom right). *Noise 9; impulse response of 30-c/s to 5-kc/s bandpass filter. 10 kc/s timing wave below.*

TABLE I

Noise Number	P.R.F. (c/s)
9	16
10	8
11	4
12	2
13	1
14	0.5
15	0.25
16	0.125
17	0.1







preceded by aural-weighting networks, these having been chosen experimentally by pilot tests not described. Table II gives the noise-meter and weighting-network combinations adopted. All noise meters were used in conjunction with preceding attenuators, the attenuation being so adjusted as to maintain constant meter reading.

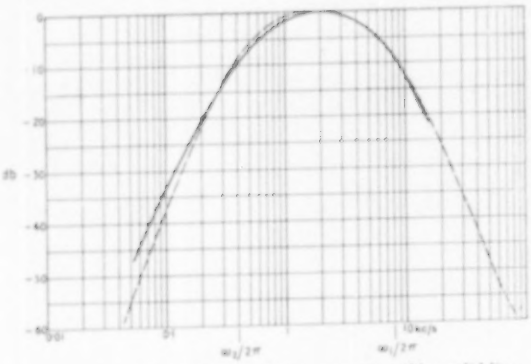


Fig. 7. *C.C.I.F. aural weighting network. The solid-line curve is the specified frequency characteristic; the dash-line is the curve of the experimental network.*

Meter (a) indicates the mean-square value of any input waveform less the mean value (usually negligible or zero). This meter will read mean-square values in a linear (undistorted) manner for all waveforms having crest factors not greater than 40 db.

Meter (b) is similar to meter (a) but is not linear for waveforms having crest factors much above 14 db.

Meter (d) indicates the full-wave rectifier mean value of a waveform applied to it.

Meters (c), (e), and (f) are all based upon the same principle, namely charging a capacitor through one resistor and discharging it through another. They are completely defined by the charge and discharge time constants if the waveform being measured is repeated with sufficient frequency for the needle of the indicating instrument to give a steady reading. This can always be assured by making both time constants large enough, and this will not affect the behaviour of the meter to varying pulse-recurrence frequencies. These meters give characteristic curves of response against p.r.f. or bandwidth (determining pulse height and width). Both these characteristic curves are unaltered if the ratio of discharge to charge times is kept constant; but they are altered if this ratio is changed. The alteration in response

TABLE II

	Noise Meter	Best Available Aural Weighting Network
(a)	B.B.C. Mean Square Meter	C.C.I.F. Aural Network for Broadcast Circuits, Fig. 7
(b)	G.P.O. Speech Voltmeter	C.C.I.F. Aural Network for Broadcast Circuits
(c)	B.B.C. Portable Noise Meter PNM/1	C.C.I.F. Aural Network for Broadcast Circuits. (This circuit is included inside the PNM/1)
(d)	V.U. Meter	7-kc/s l.p. filter in cascade with A.S.A. weighting network Z 24.3–1944 curve B
(e)	A.F. version of C.I.S.P.R. interference measuring set. Charge time :—1 ms. Discharge times switchable :—500 ms or 160 ms	5 kc/s l.p. filter
(f)	B.B.C. Peak Programme Meter PPM/6	New A.S.A. weighting network for Broadcasting Circuits, Fig. 8

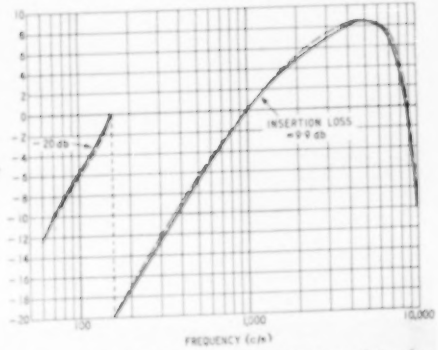


Fig. 8. *Meter (f) has the weighting curve shown by the dash-line while the solid line indicates the A.S.A. weighting curve.*

characteristic against p.r.f. for a multiplication of the ratio by a number N is a translation $1/N$ along the p.r.f. axis, Figs. 9 and 10. The actual shape of the response against p.r.f. characteristic is constant whatever the charge or discharge times and is due to the manner in which a capacitor charges and discharges. A very important practical remark results from the foregoing. It is that if a certain discharge-to-

charge ratio is found to simulate the aural annoyance against p.r.f. over a certain range of the latter, then perfectly steady meter-needle readings may be obtained, however low the p.r.f., by multiplying both charge and discharge times by a sufficiently big number. The above remarks have only been confirmed experimentally for discharge-to-charge time ratios greater than 10 and for values of p.r.f. less than a fifth of the circuit cut-off frequencies. These conditions are, however, eminently practical.

The significant characteristics of the meters for noise measurements are thus as given in Table III.

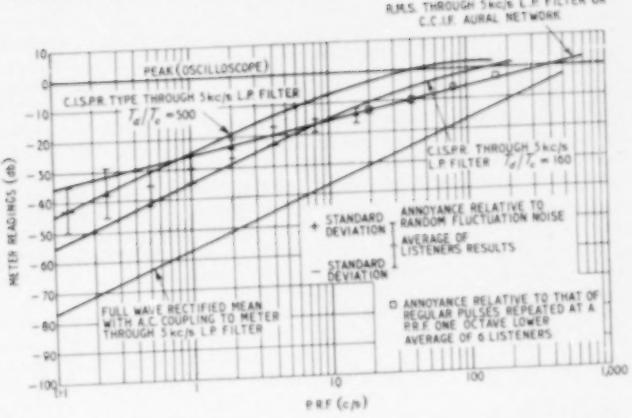

R.M.S. THROUGH 5kc/s L.P. FILTER OR C.C.I.F. AURAL NETWORK
PEAK (OSCILLOSCOPE)
C.I.S.P.R. TYPE THROUGH 5kc/s L.P. FILTER
$T_d/T_c = 500$
C.I.S.P.R. THROUGH 5kc/s L.P. FILTER $T_d/T_c = 160$
+ STANDARD DEVIATION
ANNOYANCE RELATIVE TO RANDOM FLUCTUATION NOISE
AVERAGE OF LISTENERS RESULTS
− STANDARD DEVIATION
FULL WAVE RECTIFIED MEAN WITH A.C. COUPLING TO METER THROUGH 5kc/s L.P. FILTER
□ ANNOYANCE RELATIVE TO THAT OF REGULAR PULSES REPEATED AT A P.R.F. ONE OCTAVE LOWER AVERAGE OF 6 LISTENERS
METER READINGS (db)
P.R.F. (c/s)

Fig. 9 (*above*). *Experimental results showing the meter readings with regular pulses. All meters were adjusted on random fluctuation noise applied through a 5-kc/s l.p. filter. The peak amplitude of regular impulses was made equal to the peak amplitude of random fluctuation noise, both being fed through a 5-kc/s l.p. filter.*

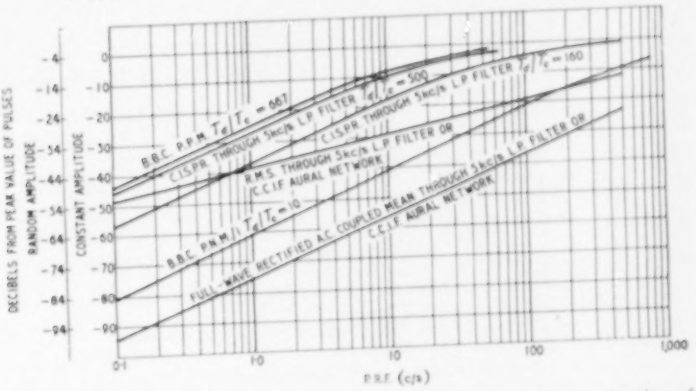

B.B.C. PPM $T_d/T_c = 667$
C.I.S.P.R. THROUGH 5kc/s L.P. FILTER $T_d/T_c = 500$
C.I.S.P.R. THROUGH 5kc/s L.P. FILTER $T_d/T_c = 160$
R.M.S. THROUGH 5kc/s L.P. FILTER OR C.C.I.F. AURAL NETWORK
B.B.C. PPM $T_d/T_c = 10$
FULL-WAVE RECTIFIED A.C. COUPLED MEAN THROUGH 5kc/s L.P. FILTER OR C.C.I.F. AURAL NETWORK
DECIBELS FROM PEAK VALUE OF PULSES
RANDOM AMPLITUDE
CONSTANT AMPLITUDE
P.R.F. (c/s)

Fig. 10 (*right*). *Calculated meter readings for regular pulses. All meters adjusted on a sine wave, taking crest and form factors into account.*

TABLE III

Noise Meter	Characteristics	
(e)	Charge time ..	.. 1 ms
	Discharge time	.. 160 ms or 500 ms
(c)	Charge time ..	.. 150 ms
	Discharge time	.. 1500 ms
(f)	Charge time ..	.. 1.5 ms
	Discharge time	.. 1000 ms

5. The Experiments

5.1 *Annoyance and Loudness as Functions of Receiver Bandwidth (Random Amplitude Pulses)*

Impulsive interference repeated at 100 c/s was fed to a v.h.f. a.m. receiver of overall bandwidth ± (30c/s to 25kc/s). An unmodulated carrier was also present. The output was passed to a loudspeaker (B.T.H. Senior R.K. with internal spider. An Altec-Lansing speaker was also tried but did not alter the laws of annoyance and loudness variation) through a de-emphasis circuit of variable time constant. Programme was fed directly to the loudspeaker from a land line. The test listener was asked to equalize by means of a calibrated attenuator the annoyance and loudness of the random amplitude output pulses emerging from the different bandwidths to that from the narrowest of them. No meter was used. The results are shewn in Fig. 11.

5.2. *Annoyance as a Function of P.R.F. (Random Amplitude Pulses)*

This experiment was similar to the previous one except that the bandwidth was fixed at 30 c/s to 5 kc/s, and the p.r.f. was varied. The test listener was asked to equalize again, with a cali-

brated attenuator, the annoyance at each p.r.f. to that obtaining when the p.r.f. was 30 c/s. The results are shown in Fig. 12.

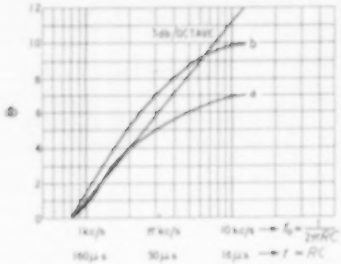


Fig. 11. *Relative annoyance and loudness levels of impulsive interference with varying bandwidth. Average curves of 7 listeners (a) no programme (loudness), (b) with programmes of speech and music (annoyance). ± 3 db deviation from mean.*

5.3. *Annoyance as a Function of P.R.F. (Constant Amplitude Pulses)*

The noises used for this test had spectra as shewn in Fig. 3(d). The test listener was asked to equalize with a calibrated attenuator the annoyance of pulses at a given p.r.f. to that obtaining for a p.r.f. one octave lower. The results are shown in Fig. 9 as □. The results of six listeners were all within ± 5 db. Again no meter was used but the results are placed upon the response characteristic of meter (a) at 10 c/s.

5.4. *Annoyance and Loudness as Read on Different Noise Meters*

The noises used for this test were those numbered one to eight. The test listener was asked to equate annoyance and loudness to that caused by the reference noise. When the test listener had equated these properties of the noises to one another and to the reference noise, the results were measured on meters (a) to (f). Tables IV, V and VI summarize the results.

Of 40 people only 3 were rejected as giving results very far from the average; 6 of the listeners were women and 30 were radio engineers. The people giving most consistent results were the least familiar with acoustic and engineering problems. Engineers with most acoustic experience gave the least consistent results. Fig. 13 shows the average change in meter readings between reference noise and each of the eight noises. Fig. 14 shows statistical frequency distributions of the test listeners' results on the various meters. Meter (a) appears to be the most

TABLE IV Loudness

Noise	1	2	3	4	5	6	7	8
No. of observations	18	18	28	19	19	18	9	9
Average of readings of meter (a) in db ..	2.81	0.805	− 0.125	− 0.105	1.6	16.6	7.71	5.83
Standard deviation in db	4.71	2.19	0.893	2.23	3.64	6.16	3.07	2.6

TABLE V Annoyance

Noise	1	2	3	4	5	6	7	8
No. of observations	28	28	43	27	26	25	14	14
Average readings of meter (a) in db ..	− 0.518	− 1.79	− 0.221	1.89	− 0.269	14.8	1.64	0.428
Standard deviation in db	3.07	2.74	1.07	2.87	4.84	6.67	2.9	2.32

TABLE VI Loudness and Annoyance

Noise	1	2	3	4	5	6	7	8
No. of observations	46	46	71	46	45	44	23	23
Average readings of meter (a) in db ..	0.32	− 0.77	− 0.18	1.06	0.523	15.2	4.03	2.54
Standard deviation in db	4.17	2.84	0.997	2.8	4.47	6.8	4.18	3.6

promising. This would seem to indicate that Ohm's Law of Hearing (D. Gabor, "New Possibilities in Speech Transmission," *J. Instn elect. Engrs*, Vol. 94, Part III, No. 32, November 1947) applies with adequate accuracy to wide spectrum type noises; that is, the ear is insensitive to small changes of phase. It is interesting to note, in this connection, that there is very little aural difference between random amplitude pulses repeated at a sufficient rate for the individual clicks to have merged into a continuous sound and random fluctuation noise. The two energy spectra can be identical, the only difference being that all pulses are in phase at zero time while random fluctuation noise may be regarded as pulses having a random statistical phase distribution. There is a difference in sound (that between frying and hissing) but it is not great.

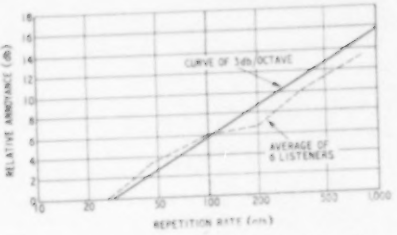


Fig. 12. *Relative annoyance of impulsive interference of random amplitude.*

5.5. *Variation of Difference in Annoyance between Random and Constant Amplitude Pulses as a Function of P.R.F.*

Noises of the type nine to twelve were used on the one hand with random amplitudes, and on the other as reference noises with constant amplitudes. The test listener was asked to adjust the level of the random-amplitude noise so that it caused the same annoyance as that produced by the same type of noise except that the amplitudes of successive pulses were constant. Table VII summarizes the results of readings of meter (a) upon the subjectively-adjusted noises.

TABLE VII Annoyance

Noise	9	10	11	12
No. of observations	14	18	9	8
Average readings of meter (a) in db.	0.822	−0.194	0.778	0
Standard Deviation in db	1.85	1.08	2.2	1.5

The meter was first 'set-up' on constant-amplitude pulses. The average change of meter reading when connected to the output of random-amplitude pulses is shown in the Table. As in 5.4 we note that meter (a) gives readings quite close to the subjective results.

5.6. *Annoyance as a Function of P.R.F. (Constant Amplitude Pulses of very Low P.R.F.)*

This test is a variation and extension of test 5.3. Instead of comparing annoyance at one p.r.f. with that caused by pulses repeated at a frequency an octave lower, the annoyance of the constant-amplitude pulses was compared with that of random fluctuation reference noise. The results are given in Table VIII.

These figures are also plotted in Fig. 9 as ×. They are plotted in such manner as to show the average test listener's ear in the form of a noise meter. Thus, for example, at a p.r.f. of 1 c/s the M.S. meter (a) would read 4½ db higher than the human noise meter, whilst the C.I.S.P.R. meter with a discharge time of 160 ms reads 5 db lower than it. The V.U. meter (d) reads 27 db lower and the C.I.S.P.R. meter with 500 ms discharge time reads 5½ db higher than it. It may be seen from Fig. 9 that at a p.r.f. below about 2 c/s the rate of fall-off of annoyance with decrease in p.r.f. appears to increase from 3 db per octave to about 6 db per octave. This would seem to shew that the ear can store energy for about a half second but not longer. It should be stated that for a p.r.f. lower than 1 or 2 c/s none of the meters used could cope with the crest factors of the impulsive waveforms present. The method adopted was to increase the p.r.f. from that listened to by the test subject to a value of 2 c/s, allowing an increase of 3 db per octave increase of p.r.f. on meter (a).

5.7. *Measurement of Programme-to-Noise Ratios.*

It seems from the foregoing that at least for the types of noises used in this series of experi-

TABLE VIII Annoyance

Noise	9	10	11	12	13	14	15	16	17
No. of observations	10	12	14	17	18	18	17	18	8
Average readings of meter (a) in db	0.625	0.25	−0.054	1.22	4.33	7.06	5.93	7.25	7.23
Standard deviation in db	2.2	1.99	2.89	2.79	4.56	6.12	8.73	7.27	7.05

ments a M.S. meter indicates annoyance adequately. Thus two major experiments remain to be done. First, is it possible to measure programme loudness with a M.S. meter? If it is, secondly, can it measure programme-to-noise ratios which have been subjectively adjusted to fit certain designations?

were then measured with meters (a) and (f), each measurement being done in a manner suitable to the particular instrument in use. Table IX shows that for the three test listeners used, the M.S. meter (a) gives a more constant reading of signal-to-noise ratio than does the peak programme meter (f).

TABLE IX

	Average r.m.s. signal-to-noise Ratio (db)	Maximum Deviation from Average r.m.s. (db)	Average signal-to-noise Ratio meter (f) (db)	Maximum Deviation from Average meter (f) (db)
Speech	30.3	1.7	32.3	1.7
Music	30	0	26.6	— 0.6

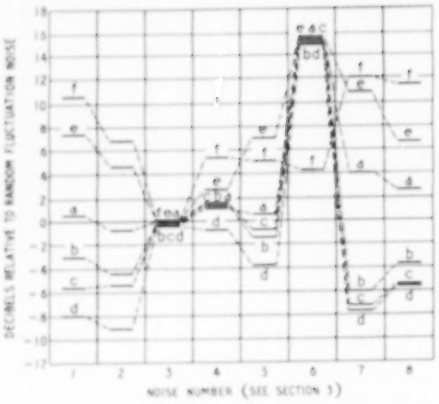


Fig. 13. Experimental results: average readings for each noise. All meters aligned on random fluctuation noise; the letters refer to the meters of Table II.

The method adopted for the measurement and comparison of programme-to-noise ratios consisted in the adjustment of a speech programme to have the same loudness as a music programme, both in the presence of a random fluctuation noise background. The signal-to-noise ratios

5.8. *Designated Programme to Noise Ratios.*

Programme was adjusted to a level of 75 phons in the absence of noise. Random fluctuation noise was then switched on and each of four test listeners was asked in turn to adjust the noise level to the description given in the first column of Table X.

TABLE X

Noise level Designation	Average r.m.s. Programme to Noise ratio (db)	Maximum Deviation from Average (db)
Just perceptible ..	60.6	3
Perceptible	50	0
Slightly disturbing ..	42	2
Disturbing	31	1

During these tests the listener could switch the programme on and off at will for cross-checking purposes. Meter (a) was used in this test.

5.9. *Effect of Noise Level on Comparison of Annoyance of Different Types of Noise.*

Test listeners were asked to adjust the levels of noises one to six so as to equate the annoyance

TABLE XI Annoyance

Noise	1	2	3	4	5	6
Number of Observations	6	6	6	6	6	6
Average increase in meter reading in db when noise level decreased by 10 db ..	3.5	2.9	0.17	2.4	8.3	— 0.42
Standard Deviation in db	4.86	4.58	1.07	4.77	9.00	6.56

to that caused by the random-fluctuation reference noise. This experiment was undertaken for two values of the reference noise level, those corresponding with 'slightly disturbing' and 'disturbing.' Table XI shews the average change in listening criterion resulted in changes of annoyance considerably less than the actual change in level. This ceases to be true for very poor signal-to-noise ratios. Meter (c) was used in this test.

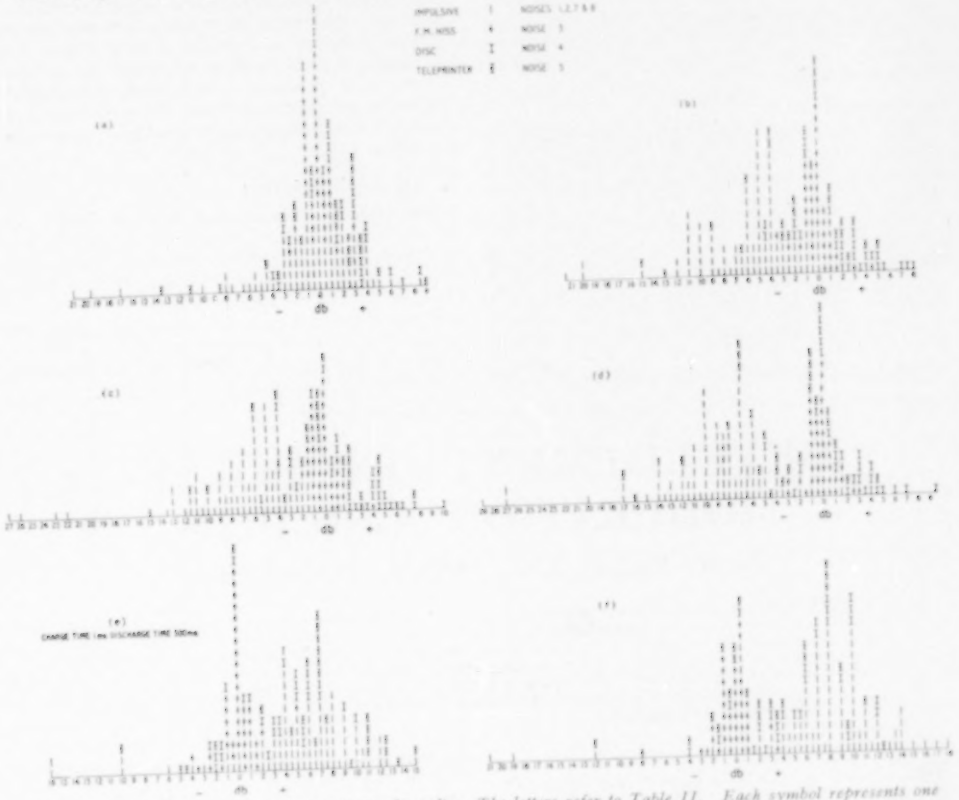


Fig. 14. *Statistical frequency distributions of results. The letters refer to Table II. Each symbol represents one observation. Abscissae show meter indications of annoyance of test noises relative to reference noise, the listener having adjusted the test and reference noises to equality of annoyance.*

(taken as positive decibels for an increase) of meter readings when the reference noise was changed from 'disturbing' to 'slightly disturbing.'

This Table would seem to indicate that the relative annoyance of impulsive noise (principally noises one, two and five) vis-à-vis random-fluctuation noise decreased slightly with decrease in level or increase in signal-to-noise ratio. The standard deviations are rather large, however, so that broadly speaking one may say that for reasonably good signal-to-noise ratios the change

TABLE XII Loudness

Noise	1	2	3
No. of observations ..	4	4	4
Average increase in meter reading in db when noise level decreased by 20 db	0.63	0.25	0
Standard deviation in db	1.08	2.28	1.37

5.10. *Effect of Noise Level on Comparison of Loudness of Different Types of Noise.*

This experiment differed from the previous one only in that programme was not used, and that a 20-db change in reference-noise level was used instead of a 10-db change. The noise levels used were 70 phons less 30 db and 70 phons less 50 db. Table XII shews virtually no change in the relative loudness of impulsive noise vis-à-vis random fluctuation noise.

6. Conclusions

The subjective effects of typical kinds of electrical interference have been studied, and a meter suitable for their measurement has been designed. The present meter, the M.S. meter (a), is, however, not suitable for p.r.f. below 1 c/s emerging from a C.C.I.F. aural network because the crest factor of such pulses would exceed 40 db and the meter would depart from true M.S. readings. It is further evident from Fig. 9 that a rate of fall of annoyance per octave decrease in p.r.f. of 6 db is required for values of p.r.f. below about 2 c/s. This characteristic could be arranged as follows: a meter of the type (a) with an improved crest factor characteristic followed by a de-emphasis circuit of time constant about half a second (to take the average of the noise waveform over a $\frac{1}{4}$-second interval) followed by a 'square-rooting' device to the output of which would be connected a d.c. indicating instrument of time constant somewhat longer than the reciprocal of the lowest p.r.f. likely to be encountered in practice. Such a device should follow the 3-db per octave law down to a p.r.f. of about 2 c/s and then change gradually to a 6-db per octave law.

7. Acknowledgments

The authors wish to express their thanks to Mr. H. Bishop, Chief Engineer of the B.B.C., for kind permission to publish this work which was undertaken at the request of Mr. H. L. Kirke, Head of the B.B.C. Research Dept.

The authors also wish to express their appreciation of the valuable advice received from Messrs. West, McMillan and Thorne of the Research Station and Radio Branch of the G.P.O.

DEVELOPMENT IN PULSE-MODULATION CIRCUITS

By F. Butler, B.Sc., M.I.E.E., M.Brit.I.R.E.

SUMMARY.—A description is given of some circuits for the generation of rectangular pulses of variable duration. They are applicable to speech transmission and a derivative of the basic circuit serves to convert the length or width modulation into pulse position or displacement modulation. The equipment is primarily intended for working over a single-channel v.h.f. radio link, but if desired, it can be arranged to provide several channels on a time-allocation multiplex basis.

The apparatus is sufficiently simple to be built into signal generators or test equipment, and the pulse parameters are readily variable.

Introduction

PULSES of high peak power are commonly produced by amplification of the output of a low-power primary generator. The process is inefficient, since certain valves in the chain must dissipate energy except during the short period in which the pulse is generated. This disadvantage has been pointed out [2,3] and a very satisfactory system has been described for the direct production of high-level pulses.[2] This circuit has been used as a basis for the present work. The following is an outline of its mode of operation:—

In Fig. 1, an alternator E maintains a sinusoidal voltage across the primary of a transformer T and develops a high voltage across the ends of the centre-tapped secondary. The frequency is that of the desired pulse-recurrence frequency. Ignoring first the impedances Z_1, Z_2, Z_3 and Z_4 and supposing the secondary output to be applied in push-pull, through the high resistances R, to the grids of the valves V_1, V_2, then during the half-cycle of applied voltage which makes the grid of one valve go negative, this valve is driven to anode current cut-off. The other valve conducts heavily, though the presence of the high grid resistance prevents dangerous over-running. This state of affairs is reversed during the next half-cycle of the driving voltage and the valve which was previously cut off now conducts. There is negligible output voltage developed across the common anode load resistance of the two valves except during the brief change-over periods when the instantaneous

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total anode current is less than its steady value.

If the phase of the grid-cathode voltage of one valve is advanced or retarded with respect to that of the other, there will be periods during which both grids are simultaneously negative and the consequent reduction in anode current produces a high-amplitude positive pulse across the load resistor. It is the function of the impedances Z_1, Z_2, Z_3 and Z_4 to introduce this small phase difference. Normally, one pair of the impedances is redundant but all four are included in the discussion for the sake of generality. Variation of the phase angle of one pair advances or retards in time the leading edge of the output pulse. Variation of the other pair of impedances similarly affects the trailing edge.

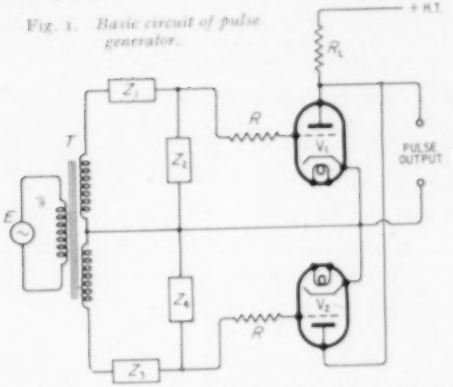

Fig. 1. *Basic circuit of pulse generator.*

Modulation of the pulse duration will be achieved if it is possible to vary the phase angle of one of the two pairs of impedances in accordance with the signal voltage. This can be accomplished by making use of the properties of a reactance modulator valve.

Reactance Modulators

A reactance modulator is essentially a thermionic valve with associated circuits so arranged that the anode-cathode impedance has a reactive component. The magnitude of the reactance is a function of the mutual conductance of the valve, which can be varied by a change in the instantaneous electrode potentials.

The two commonest reactance modulators are shown in Fig. 2.

Considering Fig. 2 (a) and ignoring the effect of the d.c. blocking capacitor in series with R and of the grid resistor in parallel with C, it is possible to calculate the anode-cathode impedance of the valve at any given frequency. At frequencies for which the reactance of C is small compared with the resistance of R, this impedance is given by :—

$$Z_i = \frac{1}{g_m} + j\omega\frac{CR}{g_m} \quad .. \quad .. \quad .. \quad (1)$$

$$\text{if } \mu \gg 1 \text{ and if } (\mu+1)\frac{1}{j\omega C} \gg R.$$

The valve thus simulates an inductance $L = C R/g_m$ in series with a resistance $1/g_m$.

The Q-factor$\left(\text{ratio}\,\frac{\text{reactance}}{\text{resistance}}\right)$ is ωCR.

Subject to the validity of the approximations made, it is interesting to note that the Q of the reactance modulator is the reciprocal of that of the associated phase-shifting circuit RC. The expression for the input impedance given in Equ. (1) does not take into account the effect of the phase-shifting circuit permanently connected in parallel with the valve. In practice, the approximation is justifiable, because the impedance of the RC circuit is very much greater than the impedance of the valve with which it is in parallel.

An exact analysis of the linear reactance modulator, taking into account the phase shifting circuit, has been given elsewhere by the writer.[6]

In the case of the circuit shown in Fig. 2 (b), the input impedance is given by :—

$$Z_i = \frac{1}{g_m} + \frac{R}{j\omega L g_m} \quad .. \quad .. \quad .. \quad (2)$$

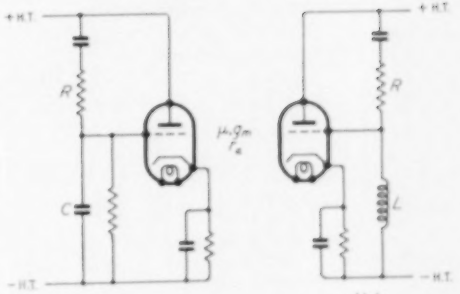

Fig. 2. *Two forms of reactance valves are shown here.*

In this case $Q = R/\omega L$ and the valve simulates a capacitance $C = g_m L/R$ in series with a resistance $1/g_m$. The series combinations of resistance and reactance given by Equs. (1) and (2) can be expressed as equivalent parallel combinations. In practice $Q < 10$ and the effect of the series resistance $1/g_m$ is equivalent to a parallel resistance Q^2/g_m, the error being about 1 per cent for $Q = 10$. To the same order of accuracy, the reactance values are unchanged

The expression for the valve reactance includes the term g_m. In the case of a triode valve, the mutual conductance is a function of the control-grid voltage. For a pentode or a hexode, g_m is also dependent on the suppressor- or mixer-grid potentials.

Referring now to Fig. 1 and assuming Z_3 and Z_4 to be replaced by reactance modulators, the possibility of pulse-length modulation is apparent. The mean value of the modulator impedance Z_1 decides the average pulse duration. Variations of this impedance under the influence of the modulation voltage cause corresponding changes in pulse duration. By proper choice of operating conditions, a linear relationship can be secured between the amplitude of the modulating signal and the mutual conductance of the reactor valve. It remains to see if this condition satisfies the requirement for proportionality between pulse duration and the amplitude of the signal voltage.

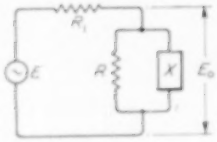

Fig. 3. *Equivalent circuit of phase-shifting network.*

Modulation Linearity

Replacing Z_1 in Fig. 1 by a pure resistance R_1 and Z_2 by the equivalent parallel circuit of a reactance modulator, the simple network shown in Fig. 3 is obtained. In this circuit,

$$\frac{E}{E_0} = \frac{R+R_1}{R} - j\frac{R_1}{X} \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

Let ϕ = phase difference between E and E_0. Then :—

$$\tan\phi = -\frac{RR_1}{R+R_1}\cdot\frac{1}{X} = -\frac{R_0}{X} \text{ where } R_0 = \frac{RR_1}{R+R_1}.$$

For a linear reactance modulator, X is inversely proportional to the mutual conductance of the valve, which, with a proper choice of operating voltages, is proportional to the instantaneous modulating signal. Short pulses are always used, so that $X \gg R_0$ and without appreciable error, $\tan\phi = \phi$. In this system of modulation, amplitude distortion is therefore small.

Variations, during modulation, of the parallel resistance of the reactor valve have been neglected. The mean value of this resistance R is, in practice, much higher than the series resistance R_1. The parallel combination of R and R_1 is, therefore, scarcely affected.

Practical Modulator Circuit

Fig. 4 is the theoretical circuit of a complete modulator with its associated pulse amplifiers, shaping circuits, audio-modulating equipment and filters.

The primary oscillator employs a pair of beam tetrode valves V_1, V_2 (Type 807) in a push-pull circuit arranged to generate a sinusoidal output of frequency 12,500 c/s. This is applied to a push-pull output transformer T with separate secondary windings. Each half-secondary supplies one grid of a 6SN7 twin-triode valve V_4. V_5 having both anodes paralleled and connected to a common load resistance R_{20}. A fixed phase-shift is applied to one of the half-secondary outputs. In the circuit of the other is included a 6SA7 valve V_3 connected as a reactance modulator, capable

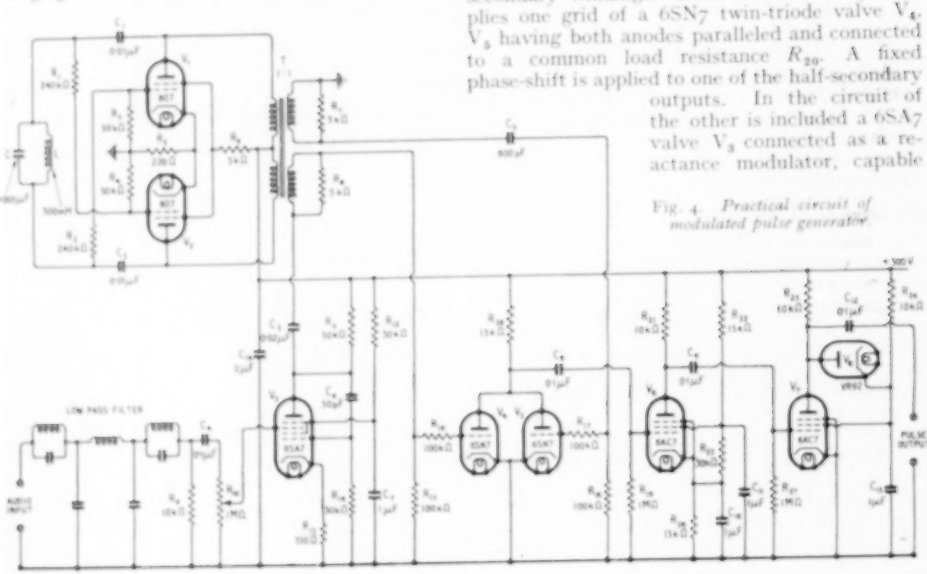

Fig. 4. *Practical circuit of modulated pulse generator.*

of simulating a variable reactance and so of causing a variable phase-displacement of the voltage developed between grid and cathode of the other half of the 6SN7 valve, referred to above. Audio-modulating voltages, applied to the grid of V_3, through a low-pass filter cutting off at 3,500 c/s, cause variations of reactor valve impedance and consequent changes in phase of the voltage derived from the primary-oscillator transformer. These phase changes cause corresponding variations in duration of the voltage pulse developed across the 6SN7 valve anode load. The remaining valves in the circuit are amplifiers and pulse-shaping stages. The output from the equipment is in the form of positive pulses, amplitude 100 volts and mean duration 1.5 microseconds.

One edge only of the output pulse is modulated, and there is thus a spurious form of pulse position modulation caused by this asymmetric operation.

Symmetrical modulation of both leading and trailing pulse edges can be achieved by the connection of a second reactor valve in series with the upper half-secondary of the primary oscillator output transformer. To achieve this, both reactor valves must be modulated by the audio signal, the polarity of which must be arranged to advance one edge of the output pulse as it retards the other.

Parallel-Tuned Phase-Shift Circuit

Examination of the impedance and phase characteristics of a parallel-tuned LC circuit shows that, near resonance, the phase angle of the impedance varies very rapidly with changes of inductance, capacitance or frequency. Such a circuit may be used as the impedance Z_2 or Z_4 in Fig. 1. The slight mistuning required to cause a phase-displacement, during modulation, can be provided by a parallel-connected reactance valve controlled by the signal voltage. Near resonance, the phase characteristic is steep and linear and a given variation of output pulse duration is achieved with a much lower audio-modulating signal than in the case of a reactance valve used alone, as in Fig. 4.

Time-Allocation Multiplex Working

The equipment shown in Fig. 4 is for single-channel working. Arrangements for two-channel operation are shown in Fig. 5.

In Fig. 5 (a) the requisite time delay between the pulse trains corresponding to Channel 1 and Channel 2 is obtained by phase-shifting bridges connected between the common primary oscillator and the two ouput transformers associated with the separate channels. A different arrangement is shown in Fig. 5 (b), where the phase-shifting circuits for interlacing the two pulse-trains are connected on the secondary side of the output transformer, coupled to a single primary oscillator.

Several other multiplex arrangements are possible. If two separate reactance valves are used, one to modulate the leading edge of the output pulse and one to modulate the trailing edge, then by feeding separate audio signals to the two reactor valves, a single pulse train may be used to carry two separate programmes.

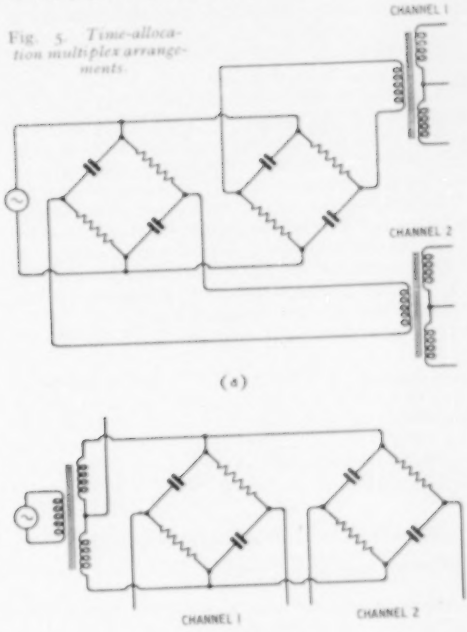

Fig. 5. *Time-allocation multiplex arrangements.*

If distortion is to be avoided in pulse-transmission systems, the highest modulation frequency is theoretically limited to one-half of the pulse-recurrence frequency. It is possible, by the employment of a band-filtering and inversion scheme, to transmit a total bandwidth equal to the pulse recurrence frequency, provided that both leading and trailing edges of the pulse train are modulated. Suppose that it is desired to transmit the frequency band 0–7 kc/s using a pulse-recurrence frequency of 7 kc/s. The sequence of operations necessary to achieve this is as follows :—

(a) Filter out the band 0–3.5 kc/s and use this to modulate the trailing edge of the pulse train as already described.

(b) Heterodyne the residue 3.5–7 kc/s against the 7-kc/s primary oscillator and filter out the sideband 0–3.5 kc/s, which will contain all the intelligence in the range 3.5–7 kc/s, although the component frequencies will have suffered inversion in the process.

(c) Use this new band of frequencies 0–3.5 kc/s to modulate the leading edge of the primary pulse train.

(d) At the receiving end, reverse these operations, separating the two channels using the technique to be described in the succeeding section. The original signal is thus recovered.

The operation is much simplified since there is available, at the receiving terminal, a train of pulses from which may be derived the precise heterodyne frequency required to reconstitute the original signal.

By an extension of the practice of band splitting, filtering and inversion, it is not difficult to evolve an elementary privacy device which, if not of high security, calls for a considerable addition to the plant at the intercepting station.

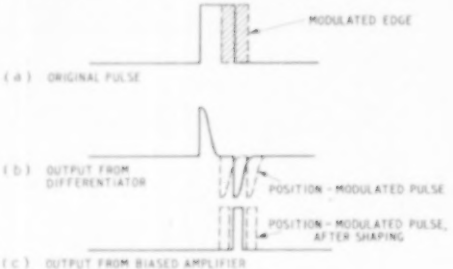

Fig. 6. *Conversion of pulse duration-modulation to position-modulation.*

It is important to realize that in all equipment used to generate pulses from sine waves, the exact definition of the start and finish of a pulse is subject to some uncertainty. If it is desired to engineer a multiplex system capable of handling some hundreds of separate channels, then the method indicated in Fig. 5 is insufficiently precise. The arrangement provides a means of generating a small number of separate pulse trains, with an adjustable time separation between each train. This facility is often useful, and the result is achieved very simply.

Displacement-Modulated Pulses

If the output of the equipment shown in Fig. 4 is passed through a short *CR* differentiator circuit, the result shown in Fig. 6 is observed.

This operation is common in radar practice, and results in the production of two exceedingly short pulses of opposite polarities from each primary pulse. The separation of the two trains is effected by biased amplifier valves.

Recurrence-Frequency Modulation

If a reactance-modulator valve is joined in parallel with one of the valves in the primary oscillator, then an audio signal applied to the reactor valve will vary the oscillator frequency and so change the recurrence frequency of the pulses. This system is not of much use in communications but has other applications.

Practical Results

The modulator circuit of Fig. 4 has been built and tested on speech transmissions. Using an audio-filter circuit cutting off at 3,500 c/s there is naturally some degradation of speech quality, but there is no noticeable difference between the signal direct from the first filter output and that after transmission, demodulation and subsequent filtering. No measurements have yet been made of the distortion of tone signals. A rough comparison of input and output waveforms at 1,000 c/s by viewing on a cathode-ray oscilloscope shows that unless the pulses are deeply modulated, such distortion is not serious, though with the circuit shown, it becomes worse at lower frequencies for some reason which has not so far been determined. Background noise is low and the 12,500-c/s tone is easily filtered out at the receiver.

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ABSTRACTS AND REFERENCES INDEX

The Index to Abstracts and References published during 1949 is in course of preparation and will, it is hoped, be available in February, priced 2s. 8d. (including postage). It includes a list of journals abstracted, together with the addresses of their editorial or publishing offices. As supplies will be limited our Publishers ask us to stress the need for early application for copies.

REFLECTION COEFFICIENT OF SNOW AND ICE AT V.H.F.

By J. A. Saxton, Ph.D., B.Sc., A.M.I.E.E.

(Communication from the National Physical Laboratory)

SUMMARY.—A review is given of the nature and composition of snow, and of the experimental knowledge of the dielectric properties of ice at very high radio frequencies. From this information an estimate has been made of the dielectric properties of snow. The presence of a layer of ice or snow on the earth's surface will produce a modification in the resultant reflection coefficient of the surface as a consequence of multiple reflections within the layer. A general formula is given in the paper by means of which the resultant reflection coefficient for plane waves in the presence of such a layer may be calculated. The formula is in terms of several basic parameters which have been determined for frequencies of 30, 300, 3,000 and 30,000 Mc/s. To illustrate the effect of these layers, the resultant reflection coefficients of a stratum of ice on sea water and of one of snow on land have been calculated for frequencies of 300 and 3,000 Mc/s, at angles of incidence of 0° (normal incidence), 45° and 80°, and for radiation polarized with the electric vector either parallel to or in the plane of incidence. It thus appears, for example, that the vertical-coverage diagram of a very high frequency radio transmitter may be appreciably modified by the presence of layers of ice or snow in depths likely to occur in practice.

1. Introduction

IN some parts of the world it is not unusual at times to have layers of snow or ice an appreciable fraction of a wavelength—or even several wavelengths—thick when the radio frequency concerned is very high, say of the order of 100 Mc/s or greater. The purpose of the present work is to calculate the effect of such layers on the reflection coefficient of the earth's surface for very high frequency radio waves. The general manner in which the problem should be tackled is obvious, it being simply a question of evaluating the effect of multiple reflections within the layer. As far as the author is aware, few such calculations have previously been made in detail, although mention should be made of the work by Pfister and Roth[1] who have considered several types of stratification including water on asphalt, ice on water and very dry sand overlying wet clay soil. The results obtained by Pfister and Roth are not readily applicable to the present problem: no investigation was made of the effects of layers of snow; furthermore a value of 5 was used for the relative permittivity of ice, a figure which is now known to be too high. In view of the possible use of very high frequencies for various purposes under conditions where ice and snow layers are likely to exist, it seems therefore that a basic knowledge of the phenomena likely to be encountered would now be desirable.

It is, of course, not possible to give in this paper a treatment of all circumstances likely to arise, since the number of independent parameters is so large, but the fundamental components of the complete solution are given together with certain selected examples to show the importance of the effects produced by ice and snow layers overlying sea water and land respectively. Any other specific case may quickly be worked out from the formulae given below for the resultant reflection coefficient and the basic data provided in Figs. 2 to 9.

It is throughout assumed that we are dealing with plane surfaces. In practice, of course, the types of surface encountered may often be rough, and when irregularities are comparable in dimensions with the wavelength we may expect the effects of multiple reflections to be somewhat blurred: even in such cases the fundamental information provided by the present calculations should be useful in the examination of snow and ice layer effects.

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2. Composition and Nature of Snow

Snow consists of ice crystals; it is, however, of a loose texture when lying on the ground, since there are considerable air spaces. Thus, to determine the effect of a layer of snow on the reflecting properties of land, besides having a knowledge of the dielectric properties of ice we require to know also the amount of ice in a given depth of snow.

Seligman[2], in a description of Alpine landscapes, says that: "When snow falls it lies on the ground in the loosest and most impalpable condition, and its specific gravity is 0.06, but under very calm conditions and at low temperatures the specific gravity may be only 0.01. The snow immediately begins to settle: the sharp points of the rays (i.e., of the crystal structure) evaporate and the flakes break up, so that the powder of broken flakes sinks together and becomes firmer. Its specific gravity will now have risen to 0.3. Then, under the influence of alternate heating and cooling a crust begins to form on

the surface, and then grows in depth. This hard crust is called 'firn snow,' it is not true ice, which is far rarer than commonly believed: in firn snow there is always some air remaining between the ice crystals." It is thus evident that the composition of a layer of snow depends much on its past history, but it would appear that generally the specific gravity of snow lying on the ground is not likely to be much greater than 0.3. This view is supported by Geddes[3] who states that on an average newly fallen snow gives a water equivalent of about one-tenth its own depth, but that snow which has lain for some time may give a yield as high as three-tenths its thickness.

3. Dielectric Properties of Ice and Snow at V.H.F.

The electrical properties of any given medium may in general be described by the relations:—

$$\epsilon = \epsilon' - j\epsilon'' = \epsilon' - 2j\sigma/f = (n - jk)^2 \quad \ldots \quad (1)$$

where
- ϵ = complex relative permittivity,
- σ = conductivity in e.s.u.,
- f = frequency in c/s,
- n = refractive index,
- k = absorption coefficient.

As is well known, the water molecule is electrically polar, so that ice, like liquid water, shows the property of anomalous dispersion at radio frequencies. Whereas this dispersion occurs in liquid water in the frequency range[4] 10^3 to 10^6 Mc/s, in ice the region of maximum dispersion is near to 6 kc/s (wavelength 50 km) at a temperature of −2° C, and occurs at even lower frequencies as the temperature decreases. The most recent observations, by Lamb[5], of the relative permittivity (ϵ') of ice at a frequency of 10,000 Mc/s (wavelength 3 cm) give a value of 3.05, and he found no variation from this value over the temperature range 0° C to −40° C. Lamb also measured the dielectric properties of ice in the frequency range 8 kc/s to 1.25 Mc/s, and his results, together with those of Smyth and Hitchcock[6], indicate that no further absorption band occurs as the frequency is increased up to 30,000 Mc/s other than that corresponding to the region of dispersion mentioned above. Consequently at frequencies of 30 Mc/s and higher, through to the centimetre waveband, the relative permittivity (ϵ') of ice may be assumed constant and equal to 3.05.

The measured[3] value of the absorption coefficient (k) of ice at 10,000 Mc/s and at 0° C is about 10^{-3}, and at −50° C it is 10^{-4}. This means that waves of this frequency will suffer an attenuation (due to absorption) of 2 db per metre of path when transmitted through ice at 0° C. According to Granier[7] the absorption coefficient of ice at −12° C and at a frequency of 10 Mc/s is 10^{-2}, which means an attenuation of 0.02 db per metre. There is no reason to suppose that the variation of k between the frequencies of 10 Mc/s and 10,000 Mc/s is not monotonic, so that the value at intermediate frequencies may be estimated if required.

As far as the author is aware, there are no direct measurements of the dielectric properties of snow, so that an attempt must be made to assess them from the known properties of ice and the density of the snow. If we assume that snow is a more or less homogeneous mixture of ice and air, it may be shown (see Appendix 1) that the effective relative permittivity (ϵ') of snow of specific gravity 0.3 is about 1.4, while for a specific gravity of 0.1 the relative permittivity may be taken as about 1.15. It may further be assumed, to sufficient accuracy, that the attenuation in snow arising from absorption in the ice is simply proportional to the amount of ice per unit volume. In this way it is found that for a specific gravity of 0.3 the attenuation is probably not greater than 0.7 db per metre and 0.007 db per metre at frequencies of 10,000 Mc/s and 10 Mc/s respectively.

In view of the relatively low absorption loss in ice and snow, and having regard to the thicknesses of ice and snow layers likely to occur in practice (which, save in exceptional circumstances, will not exceed about 1 metre) this absorption will be neglected in the following calculations of reflection coefficients. The assumption that ice and snow both behave as perfect dielectrics, with $\epsilon' = 3.05$ and $\epsilon'' = 0$ for ice, will not lead to any serious error in the values so obtained for the reflection coefficients, and it has the great advantage that from the dielectric point of view we may regard the behaviour of ice and snow as being independent of the frequency of the radiation.

An important point to be borne in mind is that if the ice or snow begins to melt, then the presence of even a small amount of liquid water will be of great importance from the point of view of absorption[4], and also to a less extent in relation to the reflectivity of a given layer. The increased absorption will be particularly significant at centimetre and decimetre wavelengths. It is difficult to see how this matter can be handled satisfactorily in a precise quantitative manner, and it must therefore be remembered that the present calculations can be expected to indicate only the order of magnitude of effects likely to occur with layers of dry ice and snow.

4. Dielectric Properties of Land and Sea

Before we can proceed to the calculation of layer effects we have still to consider what values should be assumed for the dielectric properties of soil and sea water at very high frequencies. Whereas we have assumed that for present purposes we may regard both ice and snow as pure dielectrics, a similar assumption is not possible for land and sea. The relative permittivity of soil varies over quite a large range[8,9], depending largely upon the amount of water present, and at very high frequencies usually lies somewhere between 5 and 30. The conductivity of soil also shows appreciable variation with water content, and may have values between rather less than 10^{-5} e.s.u. and about 10^{-9} e.s.u. Both components of the complex permittivity of soil, and particularly the conductivity, may be expected to decrease as the temperature of the soil falls below 0° C, but it is difficult to make a precise estimation of the magnitude of the effect in any given case. In most circumstances, however, and at any rate for frequencies not less than about 200 Mc/s, the relative permittivity of soil is several times as large as the conductivity term ($2\sigma/f$); therefore, since the relative permittivity is appreciably greater than unity whether the temperature be above or below 0° C, the reflection coefficient at the interface between air (or snow) and soil will only change slowly with changes in the relative permittivity of the soil. If an attempt were made to take account of the fact that the soil is frozen to a certain depth, it would be necessary to consider the combined effects of multiple reflections in two layers, snow and frozen soil, interposed between air and unfrozen soil. In practice the boundary between frozen and unfrozen soil may not be too well defined, and it is not likely to represent such a large discontinuity in permittivity as the snow-soil boundary: also, absorption in the soil will tend further to reduce the importance of the possible discontinuity in the soil itself. It is therefore proposed to ignore this factor: to include it would greatly increase the complexity of the calculations and, in view of the considerations given above, would be likely to lead to only a small change in the final results. In the next section formulae are given [equations (3) to (7)] for the reflection coefficient at a single plane interface separating two media, and what is required in these equations is the (complex) permittivity of the second medium relative to that of the first. Thus, suppose we have soil of conductivity (σ) 5×10^{-8} e.s.u. at a frequency of 300 Mc/s, and of permittivity (ϵ') 11 relative to vacuum (or to air with sufficient accuracy) the complex permittivity of such soil relative to air is $11 - 3.5j$. Relative to snow, where $\epsilon' = 1.4$ (relative to air) the complex permittivity of the soil becomes $(11 - 3.5j)/1.4$, or $8 - 2.5j$.

The range of frequency to be considered here is from 30 Mc/s to 30,000 Mc/s, and in arriving at reasonable probable values of the complex permittivity of soil to cover the whole of this range it is necessary to bear in mind the dispersion which occurs in water at very high radio frequencies[4]. We therefore expect that at the highest frequencies under consideration the permittivity (ϵ') of soil will decrease somewhat, while the conductivity will increase relative to the values obtaining at the lower frequencies. We therefore arrive at the complex permittivities of land relative to snow (specific gravity 0.3) given in Table I as a function of frequency. To obtain the corresponding values relative to air it is simply necessary to multiply throughout by 1.4.

TABLE I

Frequency (Mc/s)	Complex Permittivity	
	Land relative to Snow	Sea Water Relative to Ice
30	$8 - 25j$	$29 - 660j$
300	$8 - 2.5j$	$29 - 66j$
3 000	$8 - 0.8j$	$28.5 - 15j$
30 000	$4 - 0.4j$	$10 - 14j$

Table I also includes the values which have been used of the complex permittivity of sea water (at 0° C) relative to ice, and to obtain the corresponding values relative to air it is necessary to multiply throughout by 3.05. For example, the complex permittivity of sea water relative to air at a frequency of 300 Mc/s is $3.05\ (29 - 66j)$, or $88 - 200j$. The dipolar dispersion in water at very high radio frequencies is here of major importance in determining the total effective conductivity[4]. It is known that the presence of appreciable quantities of salt in water tend to reduce the permittivity (ϵ') somewhat, but for the concentration of salt occurring in sea water the effect is relatively not very great, and for the purpose of the present calculations this effect has been neglected: the inclusion of this factor would not materially change the magnitude of the calculated reflection coefficients.

5. Calculation of Reflection Coefficients

The reflection coefficients determined here are those relating to the incidence of plane waves at plane surfaces, and it is supposed that

the ice or snow layers overlying sea or land are uniform. The formulae for the reflection coefficient at a single interface between two media under such conditions were originally given by Fresnel, and they have been arranged by McPetrie[10] in a form suitable for computation in applications such as the present.

The reflection coefficient depends not only on the angle of incidence but also on the wave polarization, and there are two basic cases to be considered, one in which the electric vector of the wave is in the plane of incidence, and one in which this vector is perpendicular to the plane of incidence. Any other state of polarization may be dealt with by resolution into these two basic components.

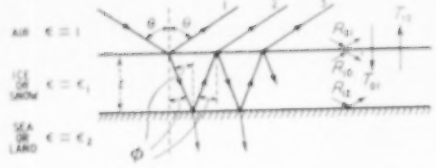


Fig. 1. *Illustration of the production of multiple reflections in ice or snow layers.*

In general a wave train suffers on reflection a change both in amplitude and phase, so that the reflection coefficient, R, is of the form :—

$$R = |R| e^{j\psi} \quad .. \quad .. \quad .. \quad (2)$$

where ψ is the advance in phase after reflection. Or, alternatively, we may write :—

$$R = K_1 + jK_2 \quad .. \quad .. \quad .. \quad (3)$$

so that $|R| = (K_1^2 + K_2^2)^{\frac{1}{2}}$ .. .. .. (4)

and $\psi = \tan^{-1}(K_2/K_1)$ .. .. (5)

For the case of a single plane discontinuity between two media McPetrie[10] has derived from the original Fresnel equations the following relations for K_1, K_2, the angle of incidence being θ :—

$$K_{1H} = \frac{\cos^2\theta - (c^2 + d^2)}{\cos^2\theta + (c^2 + d^2) + 2c\cos\theta}$$

$$K_{2H} = \frac{-2d\cos\theta}{\cos^2\theta + (c^2 + d^2) + 2c\cos\theta}$$

$$K_{1V} = \frac{(\epsilon'^2 + \epsilon''^2)\cos^2\theta - (c^2 + d^2)}{(\epsilon'^2 + \epsilon''^2)\cos^2\theta + (c^2 + d^2) + 2(\epsilon' c - \epsilon'' d)\cos\theta}$$

$$K_{2V} = \frac{-2(\epsilon' d + \epsilon'' c)\cos\theta}{(\epsilon'^2 + \epsilon''^2)\cos^2\theta + (c^2 + d^2) + 2(\epsilon' c - \epsilon'' d)\cos\theta}$$

where $c = \frac{1}{\sqrt{2}}\{[(\epsilon' - \sin^2\theta)^2 + \epsilon''^2]^{\frac{1}{2}} + (\epsilon' - \sin^2\theta)\}^{\frac{1}{2}}$

$$d = -\frac{1}{\sqrt{2}}\{[(\epsilon' - \sin^2\theta)^2 + \epsilon''^2]^{\frac{1}{2}} - (\epsilon' - \sin^2\theta)\}^{\frac{1}{2}}$$

Here the suffixes H and V refer to polarization perpendicular to and in the plane of incidence respectively. The permittivity concerned is that of the second medium relative to that of the first.

Now consider the case where we have a relatively thin layer of medium 1 interposed between the air and medium 2. Medium 1 will be either ice or snow and medium 2 either sea water or land. It will be supposed that medium 2 extends in a uniform manner indefinitely in a direction perpendicular to the interface between media 1 and 2 : thus no energy entering medium 2 returns to the air to contribute to the resultant reflected wave. The situation is as shown in Fig. 1. Since we propose to neglect the effects of absorption in medium 1, the resultant reflection coefficient, R, for an angle of incidence θ may be written (see Appendix II) as :—

$$R = |R| e^{-j\psi} = P + jQ \quad .. \quad .. \quad .. \quad (8)$$

(N.B. It is more convenient to work in terms of a retardation rather than an advance in phase.)

where :

$$P = |R_{01}|\cos\phi_{01} + \left\{\frac{A\cos y - AB\cos(z - y)}{1 + B^2 - 2B\cos z}\right\} \quad .. \quad (9)$$

$$Q = -|R_{01}|\sin\phi_{01} - \left\{\frac{A\sin y + AB\sin(z - y)}{1 + B^2 - 2B\cos z}\right\} \quad .. \quad (10)$$

$$-\psi = \tan^{-1} Q/P \quad .. \quad .. \quad .. \quad (11)$$

$$|R| = \sqrt{P^2 + Q^2} \quad .. \quad .. \quad .. \quad (12)$$

(ψ must be located in the correct quadrant by inspection of the signs of P and Q separately)

and where :

$$A = |T_{01}\,T_{10}|\,|R_{12}|; \qquad B = |R_{12}|\,|R_{10}|;$$
$$y = \delta + \psi_{12}; \qquad z = y + \psi_{10};$$
$$\delta = (4\pi t\sqrt{\epsilon_1}\cos\phi)/\lambda; \quad \sqrt{\epsilon_1} = \sin\theta/\sin\phi;$$
$$R_{01} = |R_{01}|\,e^{-j\psi_{01}}; \qquad R_{10} = |R_{10}|\,e^{-j\psi_{10}};$$
$$R_{12} = |R_{12}|\,e^{-j\psi_{12}};$$

The suffixes 0, 1, 2 refer to air and media 1 and 2 respectively. Thus R_{01} is the reflection coefficient, T_{01} the transmission coefficient and ϕ_{01} the phase retardation on reflection (6) the phase retardation on reflection in the case of a wave incident at the interface between air and medium 1, and similarly for the other symbols. δ represents the phase retardation of a wave in a single excursion in a thickness t (7) of medium 1 from the air surface to medium 2 and back to the air surface again.

Thus the basic parameters required, as a function of angle of incidence and of frequency, are R_{01}, R_{10}, R_{12} and T_{01}, T_{10}, though, for the assumptions we have made concerning the dielectric behaviour of ice and snow in the frequency range concerned, only one of these quantities, R_{12}, is a function of frequency. When the above basic parameters are known it is then possible to determine R for any angle of incidence and any thickness of medium 1—snow or ice, by substitution in equations (9) to (12).

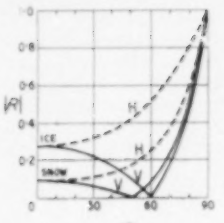


Fig. 2 (left). *Reflection coefficients of air-ice and air-snow boundaries (R_{01}) as a function of angle of incidence (θ).*

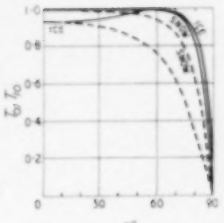


Fig. 3 (right). *Product of transmission coefficients ($T_{01} \times T_{10}$) for one transit each way across air-ice and air-snow boundaries as a function of angle of incidence on the air side (θ).*

NOTE. *Throughout Figs. 2-13 broken lines refer to polarization perpendicular to the plane of incidence (H) and full lines to polarization in the plane of incidence (V).*

6. Results and Discussion

Fig. 2 shows the variation of $|R_{01}|$ with the angle of incidence (θ) for air-ice and air-snow (specific gravity = 0.3) boundaries, and for the two basic types of wave polarization. Throughout the series of Figs. 2 to 13 the dotted lines refer to waves polarized perpendicular to the plane of incidence (horizontally polarized) and the full lines to the case of polarization in the plane of incidence (vertically polarized). Since we are assuming that ice and snow behave as perfect dielectrics (with $\epsilon' > 1$) it follows that at all angles of incidence $\psi_{01} = \pi$ for horizontally-polarized waves, and further that for vertically-polarized waves and angles of incidence less than the Brewster Angle $\psi_{01} = 0$, while when θ exceeds the Brewster Angle $\psi_{01} = \pi$. The Brewster Angle—at which $R_v = 0$—occurs at angles of incidence given by $\tan^{-1} \sqrt{\epsilon_1}$; i.e., about 50° and 60° for air-snow and air-ice boundaries respectively.

We may also determine $|R_{10}|$ from Fig. 2. To do this we remember the well-known property that :—

$$(R_{01} \text{ at angle of incidence } \theta) = -(R_{10} \text{ at angle of incidence } \phi) \quad . . \quad (13)$$

where $\sin \theta / \sin \phi = \sqrt{\epsilon_1}$

ϕ is the angle of refraction and, as is seen from Fig. 1, it is also the angle of incidence at the snow (or ice)–air boundary and the snow (or ice)–land (or sea) boundary. Thus in any specific example $|R_{01}| = |R_{10}|$. It must be noted, however, that for horizontally-polarized waves ψ_{10} is zero for all angles of incidence, and that for vertically-polarized waves when ϕ is less than the Brewster Angle $\psi_{10} = \pi$, but that when ϕ exceeds this angle $\psi_{10} = 0$; i.e., in any given case $\psi_{10} = (\psi_{01} - \pi)$. When considering reflection inside medium 1 at the boundary of medium 0 (air) it must be noted that the Brewster Angle is now given by $\tan^{-1} 1/\sqrt{\epsilon_1}$; i.e., it is about 40° and 30° for snow-air and ice-air boundaries respectively.

In order to determine $T_{01}\,T_{10}$ we make use of the further property that :—

$$T_{01}\,T_{10} = 1 - |R_{01}|^2 \quad . . \quad . . \quad (14)$$

The phase change to be associated with $T_{01}\,T_{10}$ is always zero. Fig. 3 shows $T_{01}\,T_{10}$ as a function of θ for air-snow, air-ice boundaries, and for both types of polarization.

As stated earlier, the data in Figs. 2 and 3 may be taken to cover all frequencies in the range under consideration, namely 30 to 30 000 Mc/s.

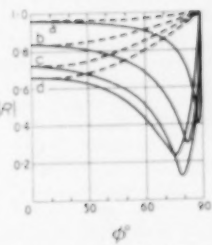


Fig. 4 (left). *Reflection coefficient of ice-sea boundary (R_{12}) as a function of angle of incidence (ϕ) and at frequencies of 30, 300, 3,000 and 30,000 Mc/s in curves a, b, c and d respectively.*

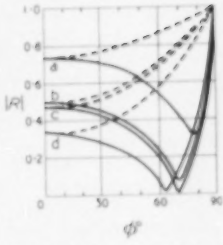


Fig. 5 (right). *Reflection coefficient of snow-land boundary (R_{12}) as a function of angle of incidence (ϕ) and at the frequencies given in Fig. 4.*

For the computation of R_{12} four frequencies have been chosen, viz: 30, 300, 3 000 and 30 000 Mc/s. Figs. 4 and 5 show $|R_{12}|$ as a function of the angle of incidence—in this case ϕ—for ice-sea and snow-land boundaries respectively, while Figs. 6 and 7 give the corresponding values of ψ_{12}. It will be observed that since ϵ'' is no longer equal to zero the Brewster Angle for vertically-polarized waves is not as well defined as when $\epsilon'' = 0$. $|R_{12}|$ never reaches zero, and

ψ_{12} changes continuously from 0 to π—although very rapidly in the neighbourhood of the pseudo Brewster Angle—and not discontinuously as in the pure dielectric case.

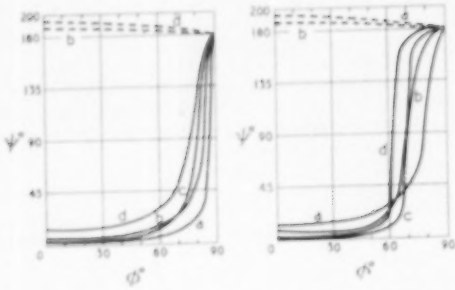


Fig. 6 (left). *Phase retardation (Ψ_{12}) on reflection at ice-sea boundary. Frequencies as in Fig. 4; for horizontally-polarized waves (Ψ_{12}) does not exceed 185° and 183° for frequencies of 3,000 and 30 Mc/s respectively.*

Fig. 7 (right). *Phase retardation (Ψ_{12}) on reflection at snow-land boundary. Frequencies as in Fig. 4; for horizontally-polarized waves (Ψ_{12}) does not exceed 183° and 184° for frequencies of 3,000 and 30,000 Mc/s respectively.*

Figs. 8 and 9 illustrate the reflection coefficient of air-sea and air-land boundaries as a function of frequency, wave polarization and angle of incidence, and enable a comparison to be made to indicate the modifications produced by ice and snow layers in any given instance.

The basic parameters provided in Figs. 2 to 7 have been used to calculate the effects of layers of snow and ice, up to one wavelength thick, overlying land and sea respectively, and for the two frequencies of 300 and 3 000 Mc/s. The calculations have been made for both types of polarization, but for three angles of incidence only: $\theta = 0$ (normal incidence, where it is unnecessary to distinguish between the two kinds of polarization), $\theta = 45°$ and $\theta = 80°$. To cover all possible cases would require a prohibitive amount of computation and number of graphs, and the above selected examples serve well to show the order of magnitude of the effects to be expected. Since it is assumed that no absorption occurs in either ice or snow the resultant reflection coefficients show a cyclic variation as a function of layer thickness, and the values for $t/\lambda > 1$ may readily be deduced from those at appropriate points within the range $0 < t/\lambda < 1$. This process must not be extended too far, however, since although absorption in the layer may be small the cyclic variation is in fact damped, and the damping may become appreciable for layer thicknesses much exceeding a wavelength, particularly in the case of ice.

Figs. 10(a), (b) and (c) show the modulus of the resultant reflection coefficient, and the phase retardation (ϕ) on reflection for a layer of ice on sea water for $\theta = 0°$, 45° and 80° respectively, and for a frequency of 3 000 Mc/s. Figs. 11(a), (b) and (c) cover similar conditions but for a frequency of 300 Mc/s. The case of air-snow-land reflection phenomena at the same angles of incidence is illustrated by Figs. 12(a), (b) and (c) for a frequency of 3 000 Mc/s, and by Figs. 13(a), (b) and (c) for a frequency of 300 Mc/s. In all the Figs. 10 to 13 short horizontal lines extending from the ordinate axis indicate what would have been the reflection coefficient in the absence of the ice or snow layer. It should be noted that the phase retardations are relative to that of the wave incident at the first boundary; i.e., the air-snow or air-ice interface.

Figs. 10 to 13 are self-explanatory, but attention might perhaps be drawn to one or two points. In the first place it will be observed that although quite large variations occur in the modulus of the resultant reflection coefficient for horizontally-polarized waves, they are such as in general to reduce $|R|$ below the value which would have obtained in the absence of the layer: only in a few instances does $|R|$ exceed the latter value, and then by a negligibly small amount.

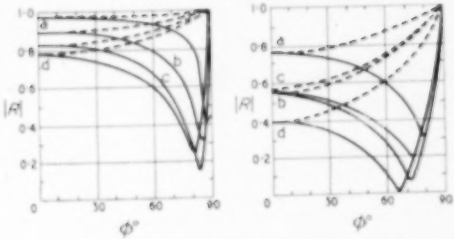


Fig. 8 (left). *Reflection coefficient of air-sea boundary. Frequencies as in Fig. 4.*

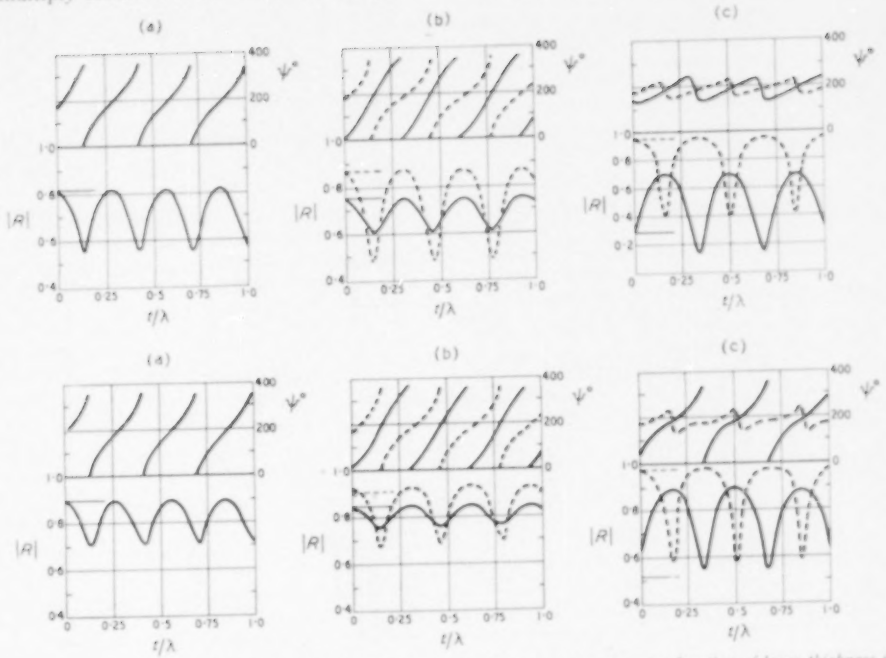
Fig. 9 (right). *Reflection coefficient of air-land boundary. Frequencies as in Fig. 4.*

In some of the examples chosen vertically-polarized waves show the same kind of variation as horizontally-polarized waves, but on the other hand we see also that with vertical polarization considerable variations in $|R|$ may occur both above and below the value obtaining in the absence of the layer. This different behaviour on the part of vertically-polarized waves is, of course, a result of the Brewster phenomenon. Thus, the angle of incidence might be such that, in the absence of an ice or snow layer, it would be near to the pseudo Brewster Angle for sea or land, and

so vertically-polarized waves would have only a quite small reflection coefficient. The interposition of a layer of ice or snow, however, materially changes the angle at which the waves are incident to the sea or land surface, and can thus give a resultant reflection coefficient considerably greater than that without the layer if the layer thickness is such as to produce a system of multiply reflected waves which are in phase

find that for an air-snow boundary $|R_v| = 0.85$ and $|R_h| = 0.89$, while for an air-ice boundary $|R_v| = 0.86$ and $|R_h| = 0.95$.

Suppose we now consider for a moment an overland transmission link in which the frequency of operation is 300 Mc/s. Let us suppose that the transmitter is at a height of at least a few wavelengths, and that the receiver is sufficiently high for the angle of incidence of waves reflected



Figs. 10 and 11. *Resultant reflection coefficient with a layer of ice on sea water as a function of layer thickness in terms of the wavelength (t/λ) ; (a) θ = 0°, (b) θ = 45°, (c) θ = 80°. Fig. 10 (top) frequency = 3,000 Mc/s. Fig. 11 (bottom) frequency = 300 Mc/s.*

when finally emerging into the air again.

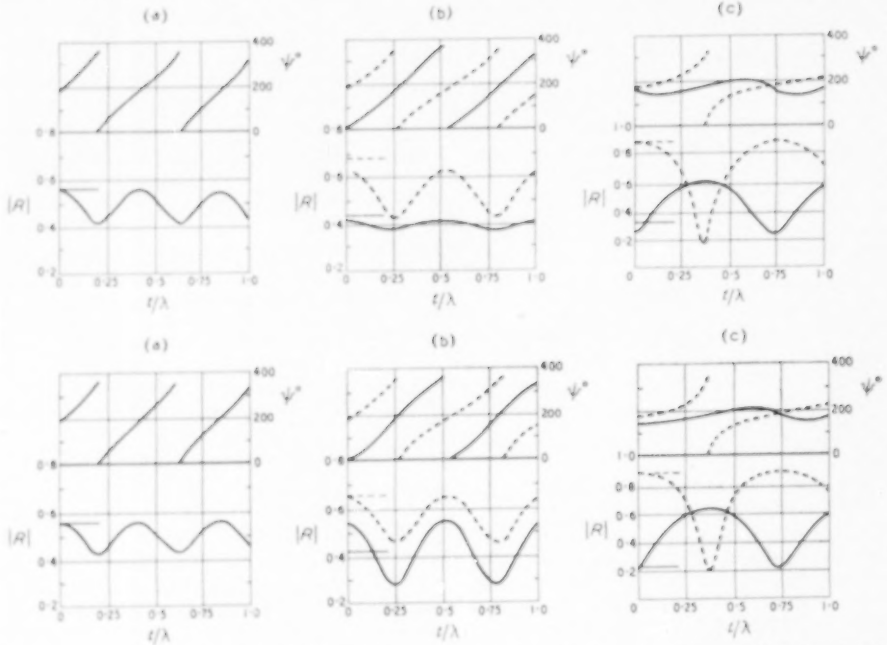
It would appear that in the case of normal incidence the reflection coefficient is likely at most to be reduced by some 20% to 35% relative to the undisturbed value ; but at higher angles of incidence, such as 45° and 80°, much bigger changes are possible. As the angle of incidence (θ) approaches 90°, however, the effects of multiple reflection must become increasingly less significant, for at grazing incidence the reflection coefficient of both air-ice and air-snow boundaries, and for both types of polarization, is unity with a phase retardation of π : no energy then enters the layer, and multiple-reflection phenomena are not possible. At $\theta = 88°$, for instance, we

at the ground to be about 80°. We further assume that the distance from the transmitter to the receiver is sufficiently large for the difference in length of the paths of the direct ray and the ground-reflected ray to be negligible, except in the determination of the phase difference between these two rays. Under such conditions the receiving terminal will be well up in the lobe structure of the polar diagram (in the vertical plane) of the transmitting aerial. Now suppose the receiver to be moved in height over an interval to embrace the nearest maximum and minimum signals ; at the above elevation this interval will be a relatively small fraction of the total height, and to a first approximation we may

assume that the reflection coefficient remains constant. In the absence of ice or snow we should expect, for horizontally-polarized waves, the maximum field strength to be of the order of $(1 + |R_n|)$, or 1.9—on some arbitrary scale—and the minimum to be $(1 - |R_n|)$, or 0.1 on the same scale. Now suppose that a layer of snow of specific gravity 0.3 exists to a depth of 0.36λ (i.e., 36 cm or 14 inches): the maximum and minimum signals will now be 1.2 and 0.8, since the resultant reflection coefficient with such a layer is only 0.2 when $\theta \approx 80°$. Compared with the situation in the absence of the snow, therefore, the maximum signal is reduced by nearly 4 db and the minimum signal increased by about 18 db. Under similar conditions vertically-polarized waves would give a maximum and a minimum of 1.25 and 0.75 respectively in the absence of snow, and of 1.65 and 0.35 with a snow layer of the above thickness (0.36λ). Thus, now, the presence of such a layer would increase the maximum signal by 2.5 db and decrease the minimum signal by 6.5 db.

An inspection of Figs. 11(a), (b) and (c) will similarly show that, whereas the reflection coefficient of sea water for horizontally-polarized waves at a frequency of 300 Mc/s, and at angles of incidence of 0°, 45° and 80°, is 0.90, 0.92 and 0.98, with a layer of ice of thickness 0.16λ (16 cm or 6.3 inches) the resultant reflection coefficients at these angles become 0.72, 0.68 and 0.58. There are, of course, variations in phase retardation, ϕ, to be considered as well, as shown in the figures. For vertically-polarized waves of this frequency, instead of the reflection coefficients of 0.90, 0.85 and 0.52 at $\theta = 0°$, 45° and 80° which occur with no layer, we obtain values of 0.72, 0.75 and 0.90 with the ice of thickness 0.16λ.

Many similar examples might be worked out, but it is apparent from the few given that the vertical-coverage diagram of a transmitter operating at very high frequencies may be appreciably modified by the presence of a layer of snow or ice, except near to the ground and at such distances that grazing incidence occurs, and when therefore $R \to -1$ in any case for both horizontally- and vertically-polarized waves.



Figs. 12 and 13. *Resultant reflection coefficient with a layer of snow on land as a function of layer thickness (t/λ); (a) $\theta = 0°$, (b) $\theta = 45°$, (c) $\theta = 80°$. Fig. 12 (top) frequency = 3,000 Mc/s. Fig. 13 (bottom) frequency = 300 Mc/s.*

7. Conclusions

The effects of multiple reflection phenomena at very high radio frequencies (300 and 3 000 Mc/s) in layers of ice and snow overlying sea water and land respectively have been investigated theoretically. The resultant reflection coefficients of the ice-sea and snow-land systems have been calculated, and it appears that marked variations may occur from the reflection coefficient which would obtain in the absence of such ice or snow layers. As a consequence the vertical coverage diagram of a very high frequency transmitter may be appreciably changed by the presence of these layers in depths likely to occur in practice frequently in some parts of the world, and occasionally in others.

8. Acknowledgments

The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of the National Physical Laboratory, and the Director of Radio Research of the Department of Scientific and Industrial Research. Mr. J. A. Lane, B.Sc., assisted with the computations.

APPENDIX I

Suppose we have a homogeneous mixture of two media having dielectric constants of ϵ_1 and ϵ_2, and that there are n_1 and n_2 molecules per c.c. of these two media respectively. Then it may be shown that, if ϵ is the resultant dielectric constant of the mixture:*

$$\left(\frac{\epsilon-1}{\epsilon+2}\right)\left\{\frac{c_1M_1+c_2M_2}{\rho}\right\}=\left(\frac{\epsilon_1-1}{\epsilon_1+2}\right)\cdot\frac{c_1M_1}{\rho_1}+\left(\frac{\epsilon_2-1}{\epsilon_2+2}\right)\cdot\frac{c_2M_2}{\rho_2} \quad \ldots \quad (15)$$

where $c_1=\dfrac{n_1}{n_1+n_2}$, $\quad c_2=\dfrac{n_2}{n_1+n_2}$

and M_1 = Molecular Weight of medium 1,

M_2 = ... 2,

ρ_1 = Density of medium 1,

ρ_2 = ... 2,

ρ = ... mixture.

To estimate the dielectric properties of snow we assume it to be a uniform mixture of air and ice. Air is itself, of course, a mixture, but for the present purpose it is sufficiently accurate to consider it as a gas of molecular weight 30 and of density 0.0013 grams/c.c. at 0°C and 76 cm pressure, and having a relative permittivity of 1.0007. The molecular weight appropriate to ice is 18, its density is 0.9 grams/c.c., and $\epsilon = 3.05$.

If now we work out ϵ_1 and ϵ_2 for snow of specific gravity 0.3 and substitute in equation (15) we find that $\epsilon = 1.4$: similarly if the specific gravity is 0.1, then $\epsilon = 1.15$.

* See, for example, "Polar Molecules" by P. Debye, 1929, The Chemical Catalog Company, Inc.

APPENDIX II

The various symbols have already been defined in the main text, Section 5. If we refer to Fig. 1 we see that the retardation in phase in ice (or snow) due to the extra path traversed in a "unit" multiple reflection is given by:—

$$\delta=\frac{4\pi t_1\sqrt{\epsilon_1}\cos\phi}{\lambda} \quad \ldots \quad (16)$$

and therefore that the total phase retardation between successive reflected rays is:—

$$(\delta+\phi_{12}+\phi_{10})=x \quad \ldots \quad (17)$$

there being no phase change to be associated with a double crossing of the air-medium 1 boundary, once in each direction.

Now the amplitude of reflected ray 1 $= |R_{01}|$

while .. .. 2 $= |T_{01}\,T_{10}\,R_{12}| = A$ (say)

and .. .. 3 $= |T_{01}\,T_{10}\,R_{12}^2\,R_{10}| = AB$ (say)

.. .. 4 $= |T_{01}\,T_{10}\,R_{12}^3\,R_{10}^2| = AB^2$ ..

.. .. etc.

thus $B = |R_{12}\,R_{10}|$

(Note.—As explained in the text we assume absorption in medium 1 to be negligible: if this were not the case it would be necessary to include an exponential attenuation factor in term B to allow for the absorption in each unit multiple reflection.)

Referring phases to that of the incident wave at the air-medium 1 boundary, we may now write the resultant reflection coefficient, R, in the form of the infinite series:

$$R=|R_{01}|e^{-j\Psi_m}+Ae^{-j(\delta+\Psi_{12})}+ABe^{-j(\delta+\Psi_{12}+x)}+AB^2e^{-j(\delta+\Psi_{12}+2x)}+\ldots \quad \ldots \quad (18)$$

i.e.,

$$R=|R_{01}|e^{-j\Psi_m}+Ae^{-jy}\{1+Be^{-jx}+B^2e^{-2jx}+B^3e^{-3jx},\ldots\} \quad \ldots \quad (19)$$

where $y=(\delta+\phi_{12})$

and, summing to infinity the geometrical progression in equation (19), we obtain:—

$$R=|R_{01}|e^{-j\Psi_m}+\left(\frac{Ae^{-jy}}{1-Be^{-jx}}\right) \quad \ldots \quad (20)$$

If now $R = P + jQ$ it is easy to show from equation (20) that:—

$$P=|R_{01}|\cos\phi_{01}+\left\{\frac{A\cos y-AB\cos(x-y)}{1+B^2-2B\cos x}\right\} \quad (21)$$

and $Q=-|R_{01}|\sin\phi_{01}-\left\{\dfrac{A\sin y+AB\sin(x-y)}{1+B^2-2B\cos x}\right\}$ (22)

Equations (21) and (22) are the same as (9) and (10) of the main text.

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RC-AMPLIFIER FILTERS

By C. H. Miller, B.E. (Tas.)

(The Physical Chemistry Laboratory, Oxford)

SUMMARY.—Simple expressions are derived for the approximate cut-off frequencies of amplifiers which use resistance-capacitance feedback networks to produce high-pass, low-pass, or band-pass frequency characteristics similar to those of conventional filters. The design of such amplifiers is discussed, and illustrations are given of the results obtained with simple amplifiers of this type.

Introduction

IT has been known for some time that at frequencies in and below the audio range an entirely RC-coupled feedback amplifier which has a high-pass, low-pass, or band-pass frequency response can have useful advantages compared with amplifiers incorporating conventional filters. In particular the omission of inductances makes for compactness, low cost, and freedom from inductive pick-up.

The feedback amplifier of Fig. 1 has a high-pass characteristic with a sharp cut-off, as has been shown by Fritzinger[1]. This can be seen by considering the vector diagram of the feedback voltages given in Fig. 2. The amplifier itself produces a phase-change of an odd multiple of 180° between its input and output terminals, and the transmission of the feedback network R_1C_1, R_2C_2 increases from zero amplitude and — 180° phase change at very high frequencies to unity amplitude and zero phase change at very low frequencies. Hence the total transmission around the feedback loop is such that there is small positive feedback at high frequencies, and large negative feedback at low frequencies.

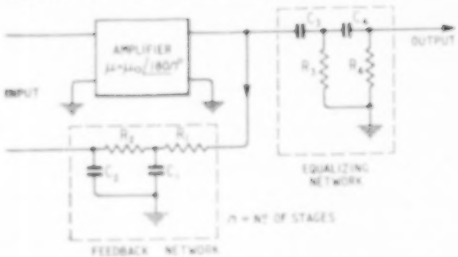


Fig. 1. *High-pass feedback amplifier.*

It follows on substituting the values of $1 - \mu\beta$ so obtained in the usual feedback amplifier expression*

$$\mu' = \mu/(1 - \mu\beta) \qquad (1)$$

that as the frequency decreases the effect of feedback is first of all to increase, and then greatly to decrease amplifier gain. Further there is one particular frequency f_0, corresponding to the point L in Fig. 2, at which the gain is a maximum and below which it rapidly decreases.

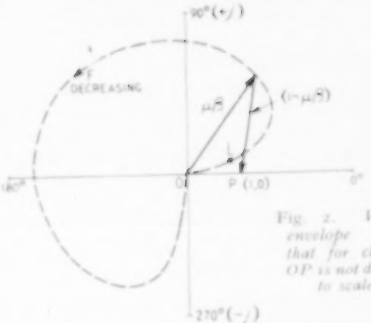


Fig. 2. *Vector envelope (note that for clarity OP is not drawn to scale).*

The resulting frequency response for an amplifier of this type is given in Fig. 3, curve (a), the response of the amplifier without feedback being very nearly flat over the frequency range considered. The amplifier may be followed by a simple equalizing network R_3C_3, R_4C_4 which has the response given by curve (b), and which makes the overall response, curve (c), of the complete amplifier, flat to within ± 1 db down to the cut-off frequency.

An amplifier with a corresponding low-pass response is obtained by interchanging R and C in each of the feedback and equalizer networks of Fig. 1, although different component values will be required for the same cut-off frequency. A band-pass response may be obtained by connecting high-pass and low-pass feedback networks in parallel, providing that their respective cut-off frequencies are sufficiently separated to prevent mutual interaction.

Cut-off Frequency

As can be seen from Fig. 3 the frequency of maximum gain, f_0, is very near the frequency of cut-off (response — 3 db), and so can be used as an approximation for it. This has the advantage that f_0 can be simply calculated, whereas a plot

MS accepted by the Editor, February 1949

[1] C. H. Fritzinger, "Frequency Discrimination by Inverse Feedback," *Proc. Inst. Radio Engrs.*, 1948, Vol. 36, p. 267.

* μ, μ' and β are, of course, both vector quantities.

of the $\mu\beta$ vector envelope would otherwise be required to obtain the cut-off frequency.

The calculation of f_0 is simplified, and remains sufficiently accurate, if the following assumptions are made. First, that the amplifier, and the networks R_1C_1 and R_2C_2, have high input and low output impedances, so that mutual loading effects can be neglected. Secondly, that over the frequency range being considered the amplifier gain is given by

$$\mu = \mu_0 \angle \frac{180n^\circ}{} \quad .. \quad .. \quad .. \quad (2)$$

where $n = 1, 3, \ldots$, the number of amplifier stages.

A further aid to calculation is to have the reciprocal *time* constants of the two feedback networks equal; i.e.,

$$\omega_1 = 1/R_1C_1 = \omega_2 = 1/R_2C_2 \quad .. \quad (3)$$

although this is not a necessary condition for correct operation.

It follows that the transmission of the feedback network is given by

$$\beta = \beta_1\beta_2 = \left\{\frac{1 - j(\omega/\omega_1)^2}{1 + (\omega/\omega_1)^2}\right\} \quad .. \quad .. \quad (4)$$

where $\omega = 2\pi f$.

On substituting from Equs. (2) and (4) we have

$$1 - \mu\beta = \frac{\{(\omega_1/\omega)^2 + 1\} + \mu_0(\omega_1/\omega)^2\{(\omega_1/\omega)^2 - 1\} - j \cdot 2\mu_0(\omega_1/\omega)^2}{\{(\omega_1/\omega)^2 + 1\}^2} \quad .. \quad .. \quad .. \quad (5)$$

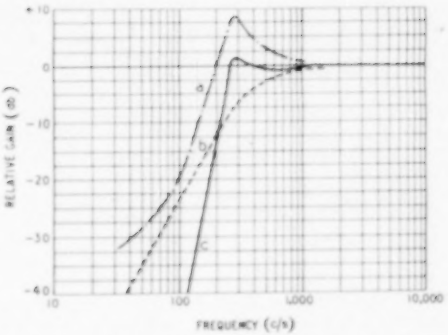


Fig. 3. *Typical high-pass amplifier response curves; (a) is for the amplifier alone, (b) is for an equalizer and (c) shows the two in combination.*

It can be seen from Equ. (1) that differentiation of $|1 - \mu\beta|^2$ with respect to ω will give the condition for the maximum value of $|\mu'|$, which is

$$(\mu_0 + 3)(\omega_1/\omega)^6 + (2\mu_0 + 5)(\omega_1/\omega)^4 + (\mu_0 + 1)(\omega_1/\omega)^2 - 1 = 0 \quad .. \quad (6)$$

For the large values of μ_0 likely to be used in practice (>100 say) this reduces with small error to

$$\omega_0 = \omega_1\sqrt{\mu_0} \quad .. \quad .. \quad .. \quad (7)$$

giving $f_0 = \dfrac{\sqrt{\mu_0}}{2\pi R_1C_1}$ .. .. .. (8)

It follows similarly for the corresponding low-pass amplifier that

$$\omega_0 = \omega_1/\sqrt{\mu_0} \quad .. \quad .. \quad .. \quad .. \quad (9)$$

$$\text{and } f_0 = 1/2\pi R_1C_1\sqrt{\mu_0} \quad .. \quad .. \quad (10)$$

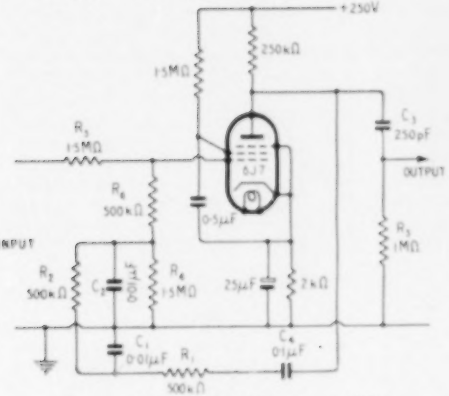


Fig. 4. *Single-stage high-pass amplifier.*

Amplifier Design

The requirements for successful design are chiefly that the assumptions made in the previous section are reasonably fulfilled. As with all feedback amplifiers care must be taken that at no time is the point P (1, 0) in Fig. 2 enclosed by the $\mu\beta$ vector envelope. This can be avoided if the condition of Equ. (2) is closely met, which is not difficult if the amplifier without feedback is designed to have a truly flat response over a sufficiently wide frequency range.

In the case of the low-pass amplifier in which f_0 is a high audio frequency care must be taken that additional phase shifts due to stray capacitance to ground are kept at a minimum. Their effect is to produce an additional lag in the feedback voltage, thus rotating the $\mu\beta$ vector away from the point P and reducing the sharpness of cut-off.

Instability can occur in a high-pass amplifier of three or more stages when the interstage RC couplings combine with the feedback network and its series d.c. blocking capacitor to act as a phase-shift oscillator at some frequency well below cut-off. This corresponds in Fig. 2 to the enclosure of the point P by the $\mu\beta$ vector from

the fourth quadrant (270°–360°). This oscillation can be avoided if the loss of the interstage couplings and the feedback network at the frequency at which they give a total phase change of 180° is sufficient to reduce the total gain around the feedback loop to less than unity.

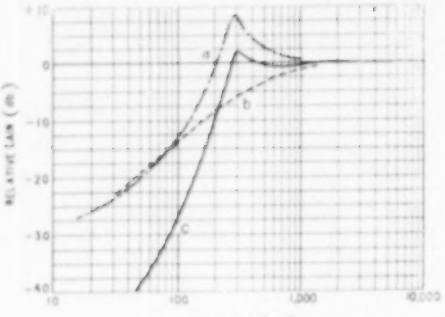


Fig. 5. Single-stage high-pass amplifier response curves; (a) and (b) respectively are for the amplifier and equaliser and curve (c) for the two together.

It is usually necessary to have the equalizing network follow the amplifier, even though this reduces the maximum output voltage available from the complete amplifier at frequencies near cut-off. Its connection directly into the amplifier input circuit is not usually possible because of the considerable phase shifts which would be introduced into the feedback loop.

The frequency response required from the equalizing network can be most speedily determined by measurement of the amplifier response after feedback has been applied. It can be seen from Fig. 2 that a small rotation of the $\mu\beta$ vector at the point L, such as is possible from unwanted phase shifts, can cause considerable variation in the value of $(1 - \mu\beta)$. For this reason the calculated increase in amplifier gain at f_0 is liable to be inaccurate, although it can be seen that such rotations of the $\mu\beta$ vector will have small effect on the accuracy of the calculated value of f_0.

Practical Amplifiers

The high-pass amplifier already referred to in Fig. 3 consists of three 6J5G triode stages which have additional negative feedback applied from unbypassed cathode-bias resistances. The gain over the feedback path has been adjusted to 54 db, and as can be seen from Fig. 3 the rate of attenuation beyond cut-off is approximately 30 db per octave.

A single RC-coupled pentode stage with a gain of the order of 40 db can provide a rate of attenuation beyond cut-off which is sufficient to be useful for many purposes. The circuit of the high-pass case of such an amplifier is given in Fig. 4, and it can be seen that for simplicity a single-section equalizing network is used. The series resistance R_5 in the grid circuit is necessary to prevent shunting of the feedback voltages by the input signal source. R_4 is used to ensure a d.c. return from grid to ground. The response curves for this amplifier are given in Fig. 5, from which it can be seen that the rate of attenuation beyond cut-off is approximately 17 db per octave. The use of a two-section equalizing network would increase this to over 20 db per octave, and would also produce a slightly flatter response down to the cut-off frequency.

The measured stage gain is found to be 77 (i.e., 37·7 db), and substituting in Equ. (8):

$$f_0 = \frac{\sqrt{77}}{2\pi \times 0.5 \times 10^6 \times 0.01 \times 10^{-6}} \doteq 280 \text{ c/s.}$$

This agrees fairly closely with the measured value of 295 c/s, as shown in Fig. 5.

In the low-pass case of this single-stage amplifier, similar rates of attenuation are obtained for cut-off frequencies of up to about 1000 c/s.

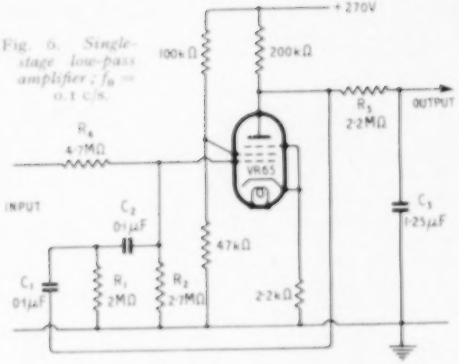


Fig. 6. Single-stage low-pass amplifier; f_0 = 0.1 c/s.

When higher cut-off frequencies are required it is necessary to use a high-slope pentode so that the same stage gain with less phase shift at high frequencies is obtained by the use of lower values of anode load resistance. Similarly it is necessary to use lower values of grid resistance to reduce phase shift in the input circuit.

The application of this type of amplifier to low frequencies is illustrated by the single-stage low-pass amplifier of Fig. 6 in which f_0 is 0.1 c/s. The resulting response curves are given in Fig. 7, which shows that the rate of attenuation beyond cut-off is approximately 20 db per octave. This particular amplifier is used to drive a pen-recording milliammeter via a cathode follower

connected to its output. In this application a very narrow bandwidth filter with a sharp cut-off is required in order to record information with the maximum signal-to-noise ratio for a given speed of response.

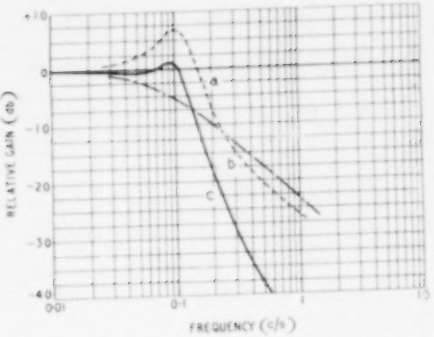


Fig. 7. *Single-stage low-pass amplifier frequency response; curves (a) and (b) are respectively for the amplifier and equalizer and curve (c) shows their combined performance.*

It is perhaps worth noting that considerable information regarding the operation of a low-pass amplifier of this type may be obtained from its response to a step voltage, or to a square wave in cases where f_0 is much over 1 c/s. As is shown in Fig. 8, the degree of overshoot indicates whether or not the peak at f_0 is correctly equalized, and also the frequency of this damped oscillation is very nearly f_0. This enables the response to be checked at frequencies for which a sine-wave generator may not be available. By this means it was checked that when C_1 and C_2 were increased to 1 μF and then to 10 μF that f_0 decreased to 0.01 c/s and 0.001 c/s respectively, as would be expected from equation (10). In the latter case equalization was obtained with R_3 increase to 28 MΩ and C_3 to 10 μF.

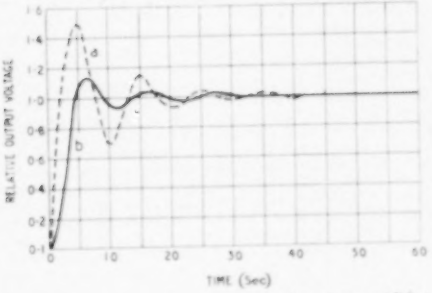


Fig. 8. *Response of a single-stage low-pass amplifier to a step voltage; curve (a) is for an amplifier with f_0 = 0.1 c/s and curve (b) shows the effect of adding an equalizer.*

As has been previously pointed out a band-pass response may be obtained by connecting low-pass and high-pass feedback networks in parallel over the same stages of an amplifier, provided that the upper and lower cut-off frequencies are sufficiently separated. Normally they have to be in a ratio of about 5 to 1. In cases where a narrower pass-band is required low-pass and high-pass amplifiers may be connected in cascade and the upper and lower cut-off frequencies adjusted independently. The response curve of a simple example of this type of amplifier is given in Fig. 9. Two RC-coupled pentode stages using VR65 valves with unbypassed cathode resistors provide an overall gain of 60 db for the two stages, including the loss in the interstage couplings. The gain around each feedback loop is approximately 40 db.

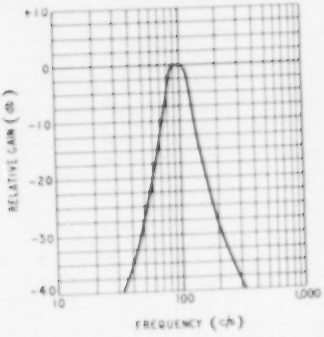


Fig. 9. *Response of a two-stage band-pass amplifier.*

Conclusion

It is evident that the advantages of this method of obtaining frequency discimination are greatest for cut-off frequencies of the order of 1000 c/s and below. At these frequencies the elimination of inductances can be a considerable advantage, and is in fact a practical necessity at frequencies below the audio range. At the higher audio frequencies there is, in many cases, little to choose between this method and the use of conventional filters as special precautions are required for optimum performance; also the inductances required for conventional filters at these frequencies are not unduly bulky or expensive.

Acknowledgment

The majority of the experimental work covered in this paper was carried out in the Electrical Engineering Laboratory of the University of Tasmania, and thanks is expressed for the assistance then obtained under the Ormsby Hamilton Research Scholarship.

BOOK REVIEWS

The Technique of Radio Design

By E. E. Zepler. Ph.D. Pp. 304 + xv, with 283 illustrations. Chapman and Hall Ltd., 37 Essex Street, London, W.C.2. Price 25s.

This is a second revised edition, the book having been first published in 1943. The author was formerly engaged in receiver design with the Telefunken Co. but is now head of the Radio Engineering Department of University College, Southampton. As stated in the preface, the book deals mainly with those problems which are closely linked with the daily routine work of an engineer in the development and testing of radio-receiving apparatus of all types. A more suitable title would have been "Radio Receiver Design." The author says that "complicated mathematics are avoided and approximations suitable to the problem in hand have been made whenever possible." In this new edition the chapter on receiver noise has been rewritten, and much more space given to negative feedback; 85 pages and 23 diagrams have been added.

Successive chapters deal with the various parts of the receiver, aerial coupling, amplifiers, detection, selectivity, noise, gain control, screening, hum, distortion, parasitic resonances, power supply, routine measurements and fault finding. Numerical examples are worked out wherever suitable, and the book is well illustrated with diagrams and graphs. Frequency modulation is only just mentioned, and the reader is referred to published papers for any further information. In the list of symbols the a.c. resistance is referred to as the impedance but is represented by R_a and not Z_a as an impedance should be. The treatment of real and reactive power on p.2 is very unsatisfactory and seems rather out of place in a book of this type. This section concludes with the following paragraph. "To think in power, either real or reactive, has often the advantage of quickly arriving at results which otherwise require an appreciable amount of calculation. It is easy to see that two different loads which, for a given E.M.F., absorb the same watts are interchangeable, watts being used in the general sense as the *sum of real and reactive power*. As the watts are the product EI the statement just made seems self-evident, because it is only another expression for the identity of two impedances. It will, notwithstanding, prove useful." We doubt it. The italics are ours.

G. W. O. H.

Cathode Ray Tube Traces

By Hilary Moss. Pp. 66 with 52 figs. and 153 photographs. "Electronic Engineering," 28, Essex St., Strand, London, W.C.2. Price 10s. 6d.

This book has an unusually large page (7¾in × 9¾in) and is set in small type, so that it contains much more material than its number of pages would lead one to suppose. The author starts by considering Lissajous figures on a geometrical basis and discusses their properties in considerable detail. In the second chapter he treats linear time bases and approximations thereto and shows that the centre part of a sinusoidal wave is adequate for many purposes. Circular and spiral time bases are discussed and there is a chapter on complex waveforms and another on modulated waves. There are appendices in which the responses of two-element series and parallel networks to various waves are discussed and in which there are useful notes on the characteristics and use of c.r. tubes. Finally, data is given on the circuits used for the production of the traces used to illustrate the book, together with some notes on linear time bases.

The photographs of c.r.-tube traces are excellent and in both their quality and their number they form an outstanding feature of the book.

W. T. C.

Die Systemtheorie der elektrischen Nachrichtenübertragung.

By K. Küpfmüller. Pp. 386 with 474 illustrations. S. Hirzel Verlag, Stuttgart.

This book is based on lectures given by Professor Küpfmüller at the Technische Hochschule in Berlin from 1937 to 1943. It is a systematic investigation of the transmission of electrical signals of all kinds under all possible conditions. It deals with telegraphic signals, speech, amplitude, frequency and pulse modulation, and all the various types of distortion. The amount of material in the book is very impressive and the treatment is very thorough and detailed and well illustrated with numerous diagrams. When any lesser known mathematical functions are introduced they are carefully explained and illustrated with numerical examples and diagrams. This is certainly a book that can be unreservedly recommended to anyone with a knowledge of German.

G. W. O. H.

Aerials for Metre and Decimetre Wavelengths

By R. A. Smith. Pp. 218 + xii. Cambridge University Press, 200, Euston Rd., London, N.W.1. Price 18s.

This book is one of a new series of monographs on Modern Radio Techniques, published under the general editorship of J. A. Ratcliffe, and dealing with the rapid advances made in the radio field during the past few years, particularly in the application of ultra-high frequencies. The aerial systems described are mainly those which have been developed for the wavelength range of 12 metres to 1 metre, and only one short chapter is devoted specifically to decimetre-wave aerials. The upper wavelength limit has been determined by two considerations: first, rigid dipoles, for the principal elements of the systems discussed, become cumbersome at greater wavelengths and are therefore not much used, and secondly 12 metres is the longest wavelength used for radar. When the wavelength falls much below 1 metre new principles of design appear. It is, of course, well known that the higher definition radar equipments, evolved during the progress of the late war were based on the use of centimetre waves; the types of aerial employed at such wavelengths, however, form the subject of another book of the series.

As Dr. Smith says in his preface: "This volume does not attempt a comprehensive description of the great variety of aerial systems used for short-wave radio applications. Indeed, most radio applications have aerial systems with their own characteristic features, so that the number of different arrangements which have been used is enormous. All that has been possible is a selection of a few applications as illustrating general principles. These illustrations have largely, though not entirely, been taken from radar applications, since it is here that perhaps the most outstanding developments in aerial design have taken place." It is, however, worth emphasizing that wavelengths in the metre-wave band have been used, and are likely to be used widely in the future, for purposes other than radar, such as in television and in radio communication links; the B.B.C. also intends to establish a high-quality broadcasting service on a wavelength of just over 3 metres (frequency 90 Mc/s).

It was already appreciated before the war, as a result of experienced gained with television, amongst other things, that just any piece of wire would not suffice as

a satisfactory receiving aerial at very short wavelengths, as had been the case to a large extent at longer wavelengths. It was also realized that accurate matching of both transmitting and receiving aerials was desirable in order to avoid spurious reflections which, with even the very short delays normally involved in the passage of the signal from aerial to receiver or transmitter, may still be of significance. The same considerations also apply in radar, where matching is also an essential feature of aerial design. The developments described in this book will, therefore, be of great interest to all workers concerned with systems operating at metre wavelengths.

One of the ways in which technique at very short wavelengths differs from that at longer wavelengths is that more efficient and highly directive aerials become possible. At centimetre wavelengths the design of aerial systems is often based on quasi-optical principles; i.e., one finds that curved metal surfaces are used as mirrors and that lenses of various types can be made to focus radio beams. Such methods are hardly practicable at metre wavelengths, although paraboloidal reflectors have found application at wavelengths at least as great as 50 cm. In general, however, in the wavelength band with which we are here concerned, aerial systems are based on various spatial arrangements of dipoles, some or all of which may be directly connected to the transmitter or receiver, and some of which may be parasitic.

The author gives a comprehensive and critical discussion of the theory underlying the behaviour of the fundamental element, the dipole. The extent to which the simple 'sinusoidal' (i.e., as regards current and voltage distributions along the length of the element) theory is valid and useful is shown, and the differences introduced by the more elaborate approaches of Hallén, Schelkunoff and King are considered. It must not be imagined that Dr. Smith is content to leave the matter on a theoretical plane for, as one who, during the war, had to design aerials for many urgent practical applications, he is careful to show how the predictions of theory fit in with such accurate measurements as have been made on typical systems. This comparison between theory and experiment is particularly important and valuable, for the theories—even the most elaborate ones—introduce certain idealizations, such as the manner of coupling the generator (or receiver) to the aerial, which it is not possible to realize in practice. It is evident that there is still room for considerable fundamental experimental work in this field.

The most common way in which aerial directivity and gain are obtained at metre wavelengths is by the use of arrays, broadside or end-fire, or of parasitic elements, directors and reflectors, as in Yagi systems. The mutual interaction between these various elements is, therefore, an important matter, and it is discussed by the author in some detail. Plane reflecting curtains, often of wire mesh, also find an application, particularly in association with broadside arrays. It is frequently necessary to support elements of an array by means of insulators, and the effect of these on the electrical lengths of such elements is described.

Other topics discussed are the special features of the design of aerials for high-power transmitters, the properties of long-wire and rhombic aerials, and the problems connected with the installation of aerials in aircraft. There is also a chapter on the design and construction of slot aerials involving the application of Babinet's principle.

In the final chapter the author discusses the way in which noise, from all sources, terrestrial and extra-terrestrial, may enter a receiver via the aerial, and summarizes his conclusions as follows: "For efficient resonant aerials sensitivity (of the receiving system) is determined by aerial noise for wavelenths greater than 5 metres. Using the best grounded-grid valves for the first stage of the receiver sensitivity is affected by aerial noise in certain directions (i.e., in relation to the Galaxy) down to wavelengths of 3 metres, but is generally determined by receiver noise. However, under conditions of abnormal sunspot activity the sensitivity may be greatly reduced at wavelengths more than a metre, particularly when directional aerials are used looking towards the sun. For shorter wavelengths sensitivity is entirely determined by receiver noise."

Dr. Smith is to be congratulated on presenting a most lucid and valuable account of the subject. The book is well produced, and liberally illustrated, and contains a wealth of practical information concerning the design of aerials. It should prove of great assistance to all who wish to construct aerials for the ever increasing applications of radio in the metre-wave band, and it may be unreservedly recommended.

J. A. S.

Radio-Frequency Heating

By L. Hartshorn, D.Sc., A.M.I.E.E., A.R.C.S., D.I.C. Pp. 237. George Allen & Unwin Ltd., Ruskin House, Museum St., London, W.C.2. Price 21s.

This book covers both induction and dielectric heating and is concerned mainly with the principles of r.f. heating. Following a short introductory chapter in which the nature of heating processes is explained there is a discussion of methods of generating r.f. power. This is fairly elementary and occupies some 30 pages only.

Induction heating is then treated in two chapters covering principles and applications. There is a detailed theoretical treatment of the production of heat in conductors of simple geometrical shape by induction and of the effect of the constants of the work coil and its coupling to the work. Localized, as well as general, heating is treated and the methods of dealing with special problems are indicated. The chapter on applications is far from being merely a resumé of common applications of r.f. heating. It runs to 60 pages and contains a great deal of useful information about materials, methods and power requirements.

Dielectric heating is similarly divided between two chapters. The principles are very thoroughly covered and under applications the author points out some of the difficulties which are encountered when attempts are made to heat inhomogeneous material. The cooking of food by r.f. heating is mentioned as an extreme example.

In discussing applications, the author deals with the advantages and disadvantages of r.f. heating in a very balanced way and gives a very fair account of the possibilities. After some of the extravagant claims which have been made for it in the past, it is refreshing to read that "It is fairly safe to say that radio-heating should not be used for any process in which the actual cost of the power is a main consideration, for alternative methods will be cheaper."

A good point about the book is that it deals mainly with just those points upon which general information is lacking—the heating coil or electrode and the material itself. The power generator is not treated in any detail for, basically it is in no way peculiar to r.f. heating. Each chapter contains an extensive bibliography and the book undoubtedly forms a useful contribution to the literature.

W.T.C.

GRAPHICAL STATISTICAL METHODS

An error occurred in this article in the December 1949 issue. In the second column of page 403 the first paragraph should read "The method of treating the data is to divide the diagram into suitable horizontal strips . . ."

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Gain of Aerial Systems

SIR,—Mr. J. Brown's letter (*Wireless Engineer*, December 1949) explains why it is difficult to make direct practical application of the 'end-fire theorem' to radiators of the waveguide type, but I fear that I must have singularly failed to make clear the more general points which arise from wire radiators.

In the first place it was known that there is a reasonably close correspondence between the gain of a broadside array of current-carrying conductors and the gain of a uniformly illuminated aperture of the same area. It is also known that if the same array of wires, which as a broadside radiator corresponded with the illuminated aperture, is fed with a different phase distribution so as to act as an end-fire radiator, the gain is doubled. Let me repeat, the *same* aerial system so far as mechanical structure is concerned, and with no question of introducing or removing an infinite plane, gives double the gain when the phasing is correctly adjusted. The general end-fire theorem was found to correspond in a number of special cases to the phasing condition which had been found to be optimum as a result of the specific calculations applicable separately to each particular case (Yagi, rhombic, polyrod, etc.).

It is possible to say that whereas the optical analogue of a broadside array of dipoles is an illuminated aperture, the optical analogue of an end-fire array is a source of the same dimensions plus a plane reflector, but I object to introducing an additional surface into the analogue when no additional component is introduced into the aerial itself.

The second question is that of an *unlimited* increase of gain, not merely 2 : 1, for an aerial system of fixed dimensions.

This is the case which inevitably conflicts with the normal ideas of optical resolution, and is one which Mr. Brown has not discussed. Subject to the limitations of conductor loss, unlimited gain should be achievable at the relatively long wavelengths where arrays of wires can be used so that drastic control of phase to fit any desired distribution function is possible.

It is not possible in the traditional type of optical device where phase is controlled by the propagation velocities through the various media employed, and the media are homogeneous over areas of a wavelength or more in diameter.

Likewise the possibility of a train of narrow pulses (*not* a single narrow pulse) through a circuit arises from analogy with the 'unlimited gain' aerial. There are functions for intensity distribution across an aperture which Fourier-transform into more peaky distributions of radiated energy versus angle than we should have thought possible before the work of authors such as Bouwkamp and de Bruijn. If those same functions are used for spectral distributions of energy, they should then Fourier-transform into amplitude versus time functions which are more peaky than we should have expected from the width of spectrum which was used.

University of Birmingham. D. A. BELL.

Phase-Shift Calculation

SIR,—The curve showing the image phase-shift coefficient β, of a multi-section filter as a function of frequency f, is often of interest in network investigations; e.g.,

1. For calculating the time-delay,

2. For calculating the effect of dissipation on the insertion loss,

3. In insertion loss calculations or in the design of Zobel filters and some insertion parameter filters.[1]

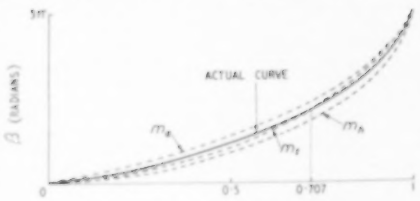
Whereas in 1 and 2 the slope of the β, f curve is the information required, in 3 the requirements are the values of f at which β is a multiple of $\pi/2$ or has some other specified value.

It is well known that the image phase-shift coefficient of a multi-section Zobel filter may be obtained as a function of frequency by adding the phase-shift coefficient curves of the individual sections. However, it is sometimes desirable to obtain a good approximation to the β, f curve more quickly.

The suggestion[2] has been made to use, as such an approximation, the phase curve of a reference filter having the same number of sections but equal m-values for all sections, the particular value of m being a suitable compromise value. It has further been suggested that the harmonic mean of the actual m-values is a suitable compromise value to be used for frequencies near the cut-off frequency(ies) (see also ref. 3).

We have examined a few compromise values and think that the results obtained may be of some general interest.

Let us, as usual in filter analysis, consider low-pass pass filters only. Let x be the normalized frequency, so that the pass band is defined by $0 \leqslant x \leqslant 1$. Let the filter consist of n half-sections with image phase-shift coefficients $\beta_\lambda/2$ and m-values m_λ. Then $\tan(\beta_\lambda/2) = m_\lambda y$ where $y = x/\sqrt{(1-x^2)}$ and the total phase-shift coefficient of the filter is $\sum_{\lambda=1}^{n} \beta_\lambda/2$. The compromise value of m will be called m_0 and the corresponding phase-shift coefficient $\beta_0/2$. Then the total phase-shift coefficient



of the compromise filter will be $n\beta_0/2$ and we are comparing $n\beta_0/2$ with $\sum\beta_\lambda/2$.

It is obvious that at $x = 0$ and $x = 1$, $n\beta_0/2$ and $\sum\beta_\lambda/2$ will be equal whatever value is chosen for m_0. However, for the cases x near to 0 and x near to 1, it can be shown (see Table I) that the best approximation of $n\beta_0/2$ to $\sum\beta_\lambda/2$ is obtained if m_0 is made equal to the arithmetic mean m_a or the harmonic mean m_h, respectively, of all the m-values.

A quantitative investigation shows that the harmonic mean, as a compromise value, gives rather large deviations from the actual β, x curve at the lower end of the range; the arithmetic mean, as a compromise value, leads to rather large deviations at the upper end of the range. In view of the fact that any value of m_0 leads to correct values of $n\beta_0/2$ at $x = 0$ and $x = 1$ (as distinct from x near to 0 and x near to 1), it seemed promising

to choose the compromise value so that a correct value of $n\beta_0/2$ would also be obtained in the interior of the range $0 \ldots x \ldots 1$, in the hope that this would give a better overall approximation to the actual curve of $\Sigma\beta_\lambda/2$. A convenient way of achieving this is to select the "tangent mean," m_t, of the m_λ terms; i.e., a value

$$m_t = \tan\left[\frac{1}{n}(\tan^{-1}m_1 + \tan^{-1}m_2 + \ldots + \tan^{-1}m_n)\right]$$

Then the curve obtained by using this compromise value $m_0 = m_t$ will cross the actual β, x curve at $x = 0.707$ (see Table I), which is near the middle of the range.

TABLE I

x	y	β_λ	$\Sigma\beta_\lambda$	$n\beta_0$
$x \to 0$	$y \to 0$	$\beta_\lambda/2 \to m_\lambda y$	$\Sigma\beta_\lambda/2 \to y\Sigma m_\lambda$	$n\beta_0/2 \to y n m_0$
$x = \frac{1}{2}\sqrt{2}$	$y = 1$	$\tan\beta_\lambda/2 = m_\lambda$	$\Sigma\beta_\lambda/2 = \Sigma\tan^{-1}m_\lambda$	$n\beta_0/2 = n\tan^{-1}m_0$
$x \to 1$	$y \to \infty$	$\pi/2 - \beta_\lambda/2 \to 1/m_\lambda y$	$n\pi/2 - \Sigma\beta_\lambda/2 \to (1/y)\,(\Sigma 1/m_\lambda)$	$n\pi/2 - n\beta_0/2 \to (1/y)\,(n/m_0)$

Two examples, both three section filters, with m-values 0.4938, 0.7454, 0.3379 and 0.5528, 1.0, 0.7806, respectively, have been investigated. The results confirm that the approximation for x near to 0 is best if $m_0 = m_s$, for x near to 1 if $m_0 = m_\infty$, and that the maximum deviations δx and $\delta\beta/2$ in the x and β directions for the whole x-range $0 \ldots x \ldots 1$ are smallest if $m_0 = m_t$. A schematic drawing of a set of curves for one filter is shown, but with exaggerated deviations because of the small scale; δx and $\delta\beta/2$ are given in Table II.

TABLE II

1st Filter			2nd Filter		
m_0	δx	$\delta\beta/2$	m_0	δx	$\delta\beta/2$
$m_s = 0.5257$	0.015	0.12	$m_s = 0.7778$	0.0175	0.07
$m_\infty = 0.4742$	0.035	0.10	$m_\infty = 0.7337$	0.020	0.06
$m_t = 0.5143$	0.009	0.06	$m_t = 0.7620$	0.005	0.03

These results suggest that m_0 should always be chosen with reference to the part of the x-range which is of chief interest.

W. Saraga.
J. Freeman.

Telephone Manufacturing Co., Ltd.
St. Mary Cray, Kent.

REFERENCES

[1] S. Darlington, Bell Monograph B-1186.
[2] This suggestion is due to Mr. H. J. Orchard (P.O. Research Station).
[3] V. Belevitch, *Electrical Communication*, June 1947.

Spontaneous Fluctuations in Double-Cathode Valves

Sir,—In a recent letter[1] K. S. Knol and G. Diemer report on the results of fluctuation measurements on electron tubes with two hot electrodes (double cathode valves) which indicate that the ratio β of "fluctuation temperature" to true electrode temperature is equal to 1 for a valve with indirectly-heated electrodes in the range 1100–1300° K and equal to about 0.9 for a valve with directly-heated electrodes in the range 2250–2550° K. The latter result is in disagreement with experiments by D. K. C. MacDonald,[2] who observed an increase of β from 1 at 1900° K to about 2 at 2450° K. In a note to *Nature*[3] I endeavoured to show that MacDonald's results could be explained on the basis of my unified theory of spontaneous electrical fluctuations,[4] and Knol and Diemer therefore feel justified in rejecting my theory. In actual fact the new experimental results neither disprove MacDonald's findings nor my theory in any way.

My theory, applied to the present case, yields the following formula for the mean square of the current fluctuation in the frequency range $f \ldots f + \Delta f$

$$(\delta I)^2_f = 4kT_{ex}\beta\Delta f \quad \ldots \quad (1)$$

where $g_x = \left(\frac{\delta I}{\delta V_x}\right)_{I_x}$ $\ldots$ (2)

is the differential tube conductivity for a fixed saturation current I_x (i.e., a fixed electrode temperature T_x) and

$$\beta = \frac{1 + \alpha^2}{1 + \alpha}, \quad \alpha = g_x/g_\epsilon; \quad \ldots \quad (3)$$

here $g_\epsilon = \epsilon\frac{eI_x}{kT}$ $\ldots$ (4)

with $\epsilon = \left(\frac{\partial I}{\partial I_x}\right)_{V_x}$ $\ldots$ (5)

The quantities g_x and g_ϵ will, in general, vary with the operating conditions of the tube (i.e., with inter-electrode voltage V_x and electrode temperature T) and they can easily be evaluated from the $I - V_x$ characteristics for a set of T.

It is seen from (3) that $\beta = 1$ for $\alpha = 0$ and $\alpha = 1$ and that it has a minimum ≈ 0.8 at $\alpha = 0.4$; for large values of α one has $\beta = \alpha$, and thus β becomes large. The critical temperature at which β starts to become greater than unity depends entirely on the valve characteristics and can be very different for tubes of different construction.

For the valve used by MacDonald I found from the characteristics that ϵ did not appreciably vary with T and for the limit $V_x \to 0$ I found by extrapolation $\epsilon_0 \approx 0.001$. With this value I deduced a critical temperature for that particular valve of about 2100° K. However, it is seen from (4) that a valve with a smaller value of ϵ_0 and the same I_x–T dependence might easily raise the critical temperature to a value well outside the range accessible to experiment. In such a case one would expect β to have a value between 0.8 and 1 which would explain the observations of Knol and Diemer with the directly-heated electrodes. For still smaller values of ϵ, β should again become equal to 1 as pointed out above.

Thus the results of Knol and Diemer, as far as they are set out in their note, cannot be used as evidence for or against my theory. MacDonald's criticism against this theory will be met elsewhere in the near future.

R. Fürth.

Birkbeck College,
University of London.

REFERENCES

[1] K. S. Knol and G. Diemer, *Wireless Engineer*, October 1949.
[2] D. K. C. MacDonald, *Wireless Engineer*, Vol. 24, p. 30, 1947.
[3] R. Fürth, *Nature*, Vol. 60, p. 832, 1947.
[4] R. Fürth, *Proc. roy. Soc.* A, Vol. 192, p. 593, 1948.
[5] D. K. C. MacDonald, *Proc. roy. Soc.* A, Vol. 195, p. 225, 1948.

Network Theorem

Sir,—The theorem discussed by E. R. Wigan in the correspondence section of December has been proved by W. Cauer[6] and applied to various problems in filter design.

Brussels. V. Belevitch.

[6] W. Cauer, "Theorie der linearen Wechselstromschaltungen", Becker & Erler, Leipzig, 1941, p. 379.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 2/- each.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

625 729.—Production of a saw-tooth deflection current, wherein the coil is shunted by a diode and a variable unbypassed resistor to improve linearity.

Marconi's Wireless Telegraph Co. Ltd. (*assignees of O. H. Schade*). *Convention date* (*U.S.A.*) *January 18th*, 1945.

625 741.—Saw-tooth generator of the capacitance charge type using a multi-electrode valve, one electrode of which is biased during the charging cycle to prevent leakage from the capacitor such as might destroy linearity.

C. L. Faudell. Application date August 13th, 1946.

626 132.—Frame synchronization is effected, in a system in which the frame synchronizing pulses are interrupted at double the line-pulse frequency, by an impedance network of specified characteristics.

Philips Lamps Ltd. Convention date (*Netherlands*) *June 15th*, 1945.

626 714.—A method of television reception using a screen consisting of transparent crystalline material which becomes opaque under radiation wherein the radiation produces discolouration dependent on scanning intensity.

Scophony Ltd. and G. Wilkenhauser. Application date August 17th, 1946. (*Addition to No.* 513 776.)

TRANSMITTING CIRCUITS AND APPARATUS

(*See also* **under** *Television*)

625 063.—Hum is reduced in mains-operated transmitters by recording the hum during non-signal conditions and playing back to suppress or reduce the hum in the output.

Philips Lamps Ltd. Convention date (*Netherlands*) *November 9th*, 1945.

625 268.—A reactance-valve frequency- or phase-modulating circuit wherein feedback is applied in respect of the modulation but not of the carrier, to secure linearity of the overall characteristic.

Philips Lamps Ltd. Convention date (*Netherlands*) *February 25th*, 1943.

625 791.—Phase-shift or modulation is effected by a multiple-electrode valve to the first grid of which an alternating input is applied while a control or modulating potential is applied to a second control grid, the output being taken from the anode.

Radio Corporation of America (*assignees of W. C. Morrison*). *Convention date* (*U.S.A.*) *June 13th*, 1946.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

625 478.—A multi-channel time-sharing communication system for moving vehicles having terminal and repeater stations wherein the received signals at the repeater stations are timed correctly to prevent interference.

Standard Telephones and Cables Ltd. (*assignees of E. M. Deloraine and D. D. Grieg*). *Convention date* (*U.S.A.*) *March 28th*, 1946.

625 623.—Selective call system for radio intercommunication using any of several frequencies in which a frequency not so used is employed to transmit a calling code and a code indicating the frequency to be used.

Standard Telephones and Cables Ltd. (*assignees of Le Materiel Telephonique, S.A.*). *Convention date* (*France*) *September 1st*, 1945.

CONSTRUCTION OF ELECTRON-DISCHARGE DEVICES

624 731.—A valve grid is formed of a platinum alloy containing 0.1% beryllium, apparently to reduce secondary emission.

The B.T.H. Co. Ltd. Convention date (*U.S.A.*) *July 3rd*, 1945.

625 259.—Velocity-modulation device in which the volume of the resonator is controlled by axial displacement of a stepped outer sleeve.

Standard Telephones and Cables Ltd. (*assignees of Le Material Telephonique, Societe Anonyme*). *Convention date* (*France*) *December 22nd*, 1943.

625 271.—A thermionic valve of the beam-deflection type with magnetic and electrical fields to cause the electrons to travel in a succession of looped paths to reduce the average speed through the valve and increase deflection sensitivity.

G. H. Metson. Application date August 1st, 1946.

625 275.—Magnetron for continuous-wave operation at high output having an anode and a coaxial cathode around it, the two being surrounded by a cylindrical housing divided into separated concentric parts.

Westinghouse Electric International Company. Convention date (*U.S.A.*) *August 28th*, 1945.

625 901.—U.h.f. amplifier valve including two aligned cavity resonators communicating by a tube and each having a secondary-emission cathode.

Compagnie Generale De Telegraphie Sans Fil. Convention date (*France*) *February 14th*, 1941.

SUBSIDIARY APPARATUS AND MATERIALS

624 767.—A closed-core inductor having a U-shaped iron-dust core and a bridge secured with a 'magnetic' adhesive (c.f. Spec. No. 616 249).

Communication Engineering Pty. Ltd. Convention date (*Australia*) *September 9th*, 1944.

625 188.—A piezo-electric crystal mounting having three metal spots around the periphery, of such thickness that when clamped between metal plates they function as energizing electrodes but are not in physical contact with the crystal.

The General Electric Company Ltd., G. M. Wells and L. Rollin. Application date March 3rd, 1947.

625 378.—Measurement of power flowing in one direction in a u.h.f. transmission line, by means of an auxiliary transmission line coupled to a rectifier and galvanometer.

Sperry Gyroscope Co. Inc. Convention date (*U.S.A.*) *August 18th*, 1943.

625 983.—A signal generator for producing 'pip' signals rich in harmonics comprises an oscillator of the desired recurrence frequency and including a highly-damped inductance having a resonant frequency much higher than the frequency of the oscillator.

W. J. O'Brien. Convention date (*U.S.A.*) *August 27th*, 1945.

ABSTRACTS and REFERENCES

Compiled by the Radio Research Board and published by arrangement with the Department of Scientific and Industrial Research

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

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References to Contemporary Papers on Acoustics.—A. Taber Jones. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 556–563.) Continuation of 2995 of 1949.

534.213.4 2

On the Propagation of Sound Waves in Narrow Conduits.—O. K. Mawardi. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 482–486.) An approximate solution of sufficient accuracy for narrow tubes of arbitrary shapes is derived and applied to the case of a wire-filled tube. Losses due to viscosity, radiation and thermal conduction are taken into account. Theoretical predictions are in satisfactory agreement with experimental results.

534.232 3

Theory of Focusing Radiators.—H. T. O'Neil. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 516–526.) The sound field due to a concave spherical radiator vibrating with uniform normal velocity is determined approximately. The radius of the circular boundary of the radiator is assumed to be large compared with λ and with the depth h of the concave surface. The point of greatest intensity is not at the centre of curvature but approaches it with increasing values of h/λ. The greatest intensity is not much greater than the intensity at the centre of curvature except when h/λ is small. The calculations are in reasonable agreement with Willard's experimental data (3003 of 1949). See also 8 below.

534.232 4

The Theory of Sound Vibrations in Open Tubes.—L. A. Weinstein. (*Zh. tekh. Fiz.*, Aug. 1949, Vol. 19, No. 8, pp. 911–930. In Russian.) The radiation of a symmetrical sound wave from the open end of a tube is discussed and a rigorous solution of the problem is given. The results obtained are also applied to the case of asymmetrical waves.

534.25 5

Refracting Sound Waves.—W. E. Kock & F. K. Harvey. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 471–481.) Refracting structures consisting of arrays of obstacles which are small compared with λ are described. These obstacles increase the effective density of the medium and thus reduce the velocity of sound passing through the array. When the dimensions of individual obstacles approach $\lambda/2$, the effective refractive index varies with λ and prisms then cause a dispersion of the waves. Path-length delay-type lenses for focusing sound waves are also described. A diverging lens is discussed which can be used in conjunction with an exponential horn loudspeaker to produce an approximately circular wave front at the higher audio frequencies. For e.m. lenses similar in principle see 2176 of 1948 (Kock). See also *Bell Lab. Rec.*, Oct. 1949, Vol. 27, No. 10, pp. 349–354.

534.321.9 6

Ultrasonics in Fluids.—E. G. Richardson. (*Nature, Lond.*, 5th Nov. 1949, Vol. 164, No. 4175, pp. 772–773.) Report of a British Association symposium. Most of the papers deal with applications of the ultrasonic interferometer.

534.321.9 : 534.833.4 7

An Optical Method for the Determination of Ultrasonic Absorption in Opaque Soft Media.—T. Hüter & R. Pohlman. (*Z. angew. Phys.*, June 1949, Vol. 1, No. 9, pp. 405–411.) Two methods are described which are particularly suitable for materials with low acoustic impedance and high absorption. The methods depend on the diffraction of light by ultrasonic waves and on the brightness distribution in the various orders of the diffraction spectrum. Results are given for various animal tissues and also for a few artificial insulating materials.

534.321.9 : 549.514.51 8

Ultrasonic Radiation from Curved Quartz Crystals.—L. Fein. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 511–516.) The electroacoustic efficiencies at 1 250 kc/s of four 1-in square x-cut quartz crystals are deduced from transmitting frequency responses in water, admittance measurements in air and water, and radiation patterns in water. One crystal is plane; the others have radii of curvature of 25, 7 and 4 cm respectively; their thicknesses are practically equal. Efficiency values deduced from acoustic measurements do not agree as well with calculated potential efficiencies as do those derived from admittance measurements;

the latter indicate that the radiation resistance is the same for the 4 crystals. The point of maximum acoustic intensity is not necessarily at the centre of curvature of the crystal. The admittance measurements in air indicate that the effective mass of a crystal decreases with increasing curvature. See also 3 above.

9

534.41 : 534.78 : 621.385.832

The Cathode-Ray Sound Spectroscope.—R. C. Mathes, A. C. Norwine & K. H. Davis. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 527–537.) A device for the rapid analysis of short samples of speech and other sounds. The energy/frequency distribution of the sound at a particular instant is displayed as a 2-dimensional pattern, and the distribution over an interval as a 3-dimensional pattern. The sample is recorded slowly on a magnetic disk, played back 200 times as fast, and analysed by a broad-band high-frequency system.

10

534.612.4

On the Reciprocity Free-Field Calibration of Microphones.—W. Wathen-Dunn. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 542–546.)

11

534.75

Some Determinants of Interaural Phase Effects.—I. J. Hirsh & F. A. Webster. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 496–501.) The binaural threshold for a pure tone presented against a background of noise depends on the interaural phase differences of the tone and the noise. Results of experiments in which a 250-c/s tone was presented against four different kinds of noise background indicate that the threshold for the tone in the presence of a periodic masking sound is not significantly dependent on interaural phase relations, but the masking and interaural phase effects increase as the frequency band of a random masking sound approaches the frequency of the tone.

12

534.78

Extraction and Portrayal of Pitch of Speech Sounds.—O. O. Gruenz, Jr. & L. O. Schott. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 487–495.) An improved method using a combination of gain control, double detection, voiced sound selection, unvoiced sound exclusion, and a means of counting the fundamental vibrations in the voiced sound intervals. Reliable indications of pitch have been obtained for frequencies in the range 100–600 c/s for a wide variety of voices. The pitch-indicating signals have been applied to a number of visual portrayal devices such as that noted in 3520 of 1946 (Riesz & Schott).

13

534.861 : 534.76

Stereophony.—K. de Boer. (*Tijdschr. ned. Radiogenoot.*, Sept. 1949, Vol. 14, No. 5, pp. 137–146.) General discussion of the basic principles of recording and of reproduction for an audience.

14

621.392.51

Theoretical Aspects of the Reciprocity Calibration of Electromechanical Transducers.—S. P. Thompson. (*J. acoust. Soc. Amer.*, Sept. 1949, Vol. 21, No. 5, pp. 538–542.) A unified theory is presented in a form from which particular calibration theories can be derived. Two types of calibration procedure are discussed as special cases.

15

621.395.61

Directional Microphone.—H. F. Olson & J. Preston. (*RCA Rev.*, Sept. 1949, Vol. 10, No. 3, pp. 339–347.) The development of a second-order gradient system, with ribbon-type units, is described. The system has a uniform and narrow directivity pattern, and a smooth frequency-response characteristic between 50 and 15 000 c/s. In conventional studios, speech can be picked up at a distance of 12 ft. By using a monitoring console in conjunction with several of these microphones in fixed positions, so that each covers a section of the studio area, large and rapidly changing areas of action can be covered with less variation in output level than is possible with the conventional microphone and boom arrangement.

16

621.395.623.7

Third-Class Loudspeakers with Low Power Consumption.—P. V. Anan'ev. (*Radiotekhnika, Moscow*, July/Aug. 1949, Vol. 4, No. 4, pp. 18–27. In Russian.) Cheap loudspeakers, which would be effective in operation and whose construction would not require materials in short supply, are needed for remote collective farms. Loudspeakers using Rochelle salt crystals appear to meet the requirements. Such loudspeakers, provided with a volume control and a special correcting circuit, are described in detail; many performance curves are included.

17

621.395.623.7

New 15-Inch Duo-Cone Loudspeaker.—H. F. Olson, J. Preston & D. H. Cunningham. (*Audio Engng, N.Y.*, Oct. 1949, Vol. 33, No. 10, pp. 20–22, 48.) Modifications of the R.C.A. loudspeaker Type LC1A which make mass production possible are discussed. See also 993 of 2664 of 1947 (Olson & Preston).

18

621.395.623.8

Ground Loudspeakers.—D. Scott. (*Audio Engng, N.Y.*, Oct. 1949, Vol. 33, No. 10, pp. 18–19.) See 19 below.

19

621.395.623.8

Underground Loudspeakers.—J. Merhaut. (*Tesla tech. Rep., Prague*, March 1949, pp. 35–38.) Description of the loudspeakers set level with the ground in the Strahof stadium at Prague for use in connection with mass gymnastic and other displays in an arena 1 000 ft by 650 ft. The diaphragm is made of synthetic humidity-resisting material. The horn is of the folded exponential type. The complete unit is sealed and thus protected against humidity. Performance is unaffected even by rainwater. Arrangements are provided for easy maintenance or replacement of the loudspeaker unit without removal of the main cover, which is strong enough for a lorry to be driven over it.

20

621.395.625.2

Determination of the [stylus] Velocity Amplitude in 'Constant-Velocity' Recording, by Measurement of the Width of a Luminous Band.—J. H. (*Radio tech. Dig., Édn franç.*, Oct. 1949, Vol. 3, No. 5, pp. 303–305.) Short description of the method of Buchmann & Meyer, in which the disk groove is illuminated by a parallel beam of light and the pattern formed by the reflected beam on a screen a few metres away is examined.

21

621.395.625.3

New Type of Magnetic Recorder.—L. A. Fishoff. (*Radio tech. Dig., Édn franç.*, Oct. 1949, Vol. 3, No. 5, pp. 291–294.) Short description of the principal features of a tape recorder in which tape friction is avoided by using magnetic coupling. Some details of the mechanism, the amplifier and the motors are included.

22

621.395.625.3

Magnetic Recording.—M. Alixant. (*Radio tech. Dig., Édn franç.*, Oct. 1949, Vol. 3, No. 5, pp. 259–291. Bibliography, pp. 294–301.) Basic principles and modern tapes and recording heads are discussed. Details are given of many commercial instruments for recording on wire or tape. Most of these are of American manufacture, but some French, Swiss, German and English recorders are included.

621.395.625.3 : 621.395.813 23

Techniques for Improved Magnetic Recording.—L. C. Holmes. (*Elect. Engng., N.Y.*, Oct. 1949, Vol. 68, No. 10, pp. 836–841.) Essential substance of a 1948 National Electronics Conference paper. Summary noted in 2423 of 1949.

621.395.92 24

A Modern Hearing-Aid.—(*Radio Tech., Vienna*, Oct. 1949, Vol. 25, No. 10, pp. 613–615, 622.) Description, with detailed circuit diagram, of the 'Vienna' light-weight equipment using subminiature components. The frequency-response curve can be adjusted and a.g.c. is provided. The crystal microphone, amplifier and batteries are enclosed in a flat moulded case 12 × 6.5 × 2.5 cm which can easily be carried inside a coat or in a pocket. Four models are available, suitable for different degrees of deafness. One model is of a simpler type without tone adjustment.

621.395.625 25

Elements of Sound Recording. [Book Review]—J. G. Frayne & H. Wolfe. Publishers: J. Wiley & Sons, New York, 1949, 686 pp., $8.50. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 233–234.) The first 12 chapters review fundamental material on microphones, amplifiers, attenuators, equalizers etc. The main interest of the authors is in sound *film* recording. "... a very welcome addition to the literature of sound recording."

AERIALS AND TRANSMISSION LINES

621.315.2 26

Highly Balanced Radio-Frequency Transmission Lines.—K. H. Zimmermann. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, pp. 201–203.) The twinax line differs from a coaxial line in having two identical inner conductors which are insulated separately and then twisted. The same dielectric material is extruded over the inner conductors after twisting. The braided shield and thermoplastic jacket are then applied as for a coaxial line. Formulae and graphs are given for the characteristic impedance of twinax lines; the practical determination of various parameters is briefly discussed.

621.315.212 27

Coaxial Cable with Confocal Elliptical Cylindrical Conductors.—M. R. Shebes. (*Radiotekhnika, Moscow*, July/Aug. 1949, Vol. 4, No. 4, pp. 36–44. In Russian.) The effects of deformation of the cylindrical conductors of a coaxial cable on its parameters are discussed. The results obtained for the characteristic impedance and attenuation are compared with those obtained for cables with (*a*) non-concentric circular cylindrical conductors, (*b*) cylindrical conductors whose cross-sections are bounded by limaçons.

621.392.1 28

Terminal Impedance and Generalized Two-Wire-Line Theory.—R. King & K. Tomiyasu. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1134–1139.) Conventional transmission-line theory neglects variations in the parameters of the line, and coupling near a terminating impedor or near any other departure from uniformity. A theory is derived in which a simple terminal-zone network N of lumped series and shunt elements is introduced to allow for these effects. Such a network can be determined for each type of termination or discontinuity. The apparent terminal impedance Z_{ss}, which is the impedance actually measured on a lossless line at a distance $\lambda/2$ from the termination, is the impedance of the network formed by combining N with the theoretical isolated impedance Z_s of the load. For a fixed termination and given Z_s, Z_{ss} may vary greatly with the nature of the connection to the line, the relative orientation of line and load, and the type of line and its dimensions.

621.392.1 29

Energy Relationships for a High Frequency Transmission Line.—A. L. Fel'dshteyn. (*Radiotekhnika, Moscow*, July/Aug. 1949, Vol. 4, No. 4, pp. 45–50. In Russian.) Discussion taking account of losses and specified terminations. Conditions for optimum energy transfer are established and a conception of Q for the combination of oscillator, line and load is introduced.

621.392.26† 30

Spatial Beating in Coupled Waveguides.—P. E. Krasnushkin & R. V. Khokhlov. (*Zh. tekh. Fiz.*, Aug. 1949, Vol. 19, No. 8, pp. 931–942. In Russian.) Spatial beating can be observed in two parallel coupled waveguides in each of which symmetrical waves are propagated. The phenomenon consists of a periodic transference of wave energy from one waveguide to the other. The phenomenon can be regarded as an analogy in space of the beating which takes place between two coupled oscillatory systems. The case of two semi-elliptical waveguides coupled through a slot is here investigated theoretically and experimental results are given.

621.392.26† : 517.54 31

Application of Conformal Representation to the Field Equations for Rectangular Waveguides of Non-Uniform Cross-Section.—R. Piloty, Jr. (*Z. angew. Phys.*, Aug. 1949, Vol. 1, No. 19, pp. 441–448.) By means of conformal transformation, the field at an irregularity in either the E-plane or the H-plane of a rectangular waveguide can be referred to the field in a plane-walled guide filled with a medium whose permeability or dielectric constant varies from point to point. For excitation by H_{10} waves, a relatively simple partial differential equation for the field can be derived. It is also shown that by means of a simple frequency transformation the field at an E-irregularity in a rectangular waveguide can be referred to that at a corresponding irregularity in a band-pass transmission line. A method of solving the general equation, which among other things enables the impedance transformation of the waveguide quadripole to be determined, will be given in a later paper. See also 2427 and 2428 of 1949 (Rice).

621.392.26† : 621.392.5 32

Basis of the Application of Network Equations to Waveguide Problems.—Kerns. (*See* 56.)

621.392.3.012.1 33

A Simple Vector Diagram for High-Frequency Lines.—P. Cornelius. (*Commun. News*, June 1949, Vol. 10, No. 2, pp. 33–40.) The input impedance of a lossless h.f. transmission line can be determined, for a given terminating impedance, by means of a Smith chart; the attenuation is only given if the line is terminated by its characteristic impedance. The impedance of a line with losses, terminated arbitrarily, can be determined by the vector method here given, but the attenuation is only given for termination by the characteristic impedance. The vector method is also used to derive various well-known relations.

621.392.43 34

Two-Band Antenna-Matching Networks.—J. G. Marshall. (*QST*, Oct. & Nov. 1949, Vol. 33, Nos. 10 & 11, pp. 14–18 & 48–51., 114.) Continuation of 3869 of 1945. All practical cases of aerials working on two harmonically related amateur bands are discussed, with straightforward design formulae.

621.396.67 35

Short-Antenna Characteristics — Theoretical.—L. C. Smeby. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1185–1194.) A mathematical analysis

of experimental data obtained by Smith & Johnson (651 of 1948) for aerials with umbrella top-loading. The theory developed agrees satisfactorily with the experimental results. For single-tower operation, the horizontally polarized radiation is negligible. With the optimum length of umbrella, the vertical radiation characteristic is the same as it would be from the radiator without top loading. The method of analysis can be extended to other kinds of top loading.

621.396.67 : 621.397.6 **36**

"Supergain" TV Antenna developed by RCA.—(*Broadcast News*, Sept. 1949, No. 56, pp. 4–6.) An aerial consisting of dipoles mounted in front of reflecting screens $\frac{1}{4}\lambda$ wide and $\frac{1}{4}\lambda$ high fixed on the four sides of a tower. The screens are connected electrically to the tower and to each other at their vertical edges. Various methods of connection to obtain directional patterns of different shapes are possible; rectangular dipole shielding wings connected to the edges of the screens at an angle of 135° can be used to prevent undesired mutual coupling of the dipoles. The aerial is usually fed by a single transmission line. Radiators are spaced 0.9λ apart vertically, so that current distribution is nearly uniform. The power gain relative to a dipole, averaged over 360° azimuth, is $1.1n$, where n is the number of bays of dipoles.

621.396.67 : 621.397.6 **37**

A Tunable Built-In TV Antenna.—R. B. Albright. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 134–150.) This Philco aerial is a $\lambda/2$ dipole consisting of two tapered sections of Al foil 0.005 in. thick. Signals can be received efficiently from all the 12 existing U.S. v.h.f. television channels. A tunable matching circuit, which includes a variable capacitor and three fixed inductors, enables the user to tune in to each channel so as to obtain the best response and to eliminate interference. The equivalent circuit of the complete system and resonance conditions are discussed for the high and the low television frequency bands. See also *Tele-Tech*, Oct. 1949, Vol. 8, No. 10, pp. 37, 60.

621.396.67.029.54 **38**

Mode of Action and Design of Modern Broadcasting Transmitting Aerials for the Medium-Wave Range.—K. Fischer. (*Elektrotech. u. Maschinenb.*, Sept. 1949, Vol. 66, No. 9, pp. 237–242.) Discussion of ground-wave and space-wave effects, radiation diagrams of ordinary aerials, and special aerials designed to minimize fading.

621.396.671 **39**

Vertical-Aerial Radiation Characteristics over Uneven Terrain.—H. Köhler. (*Elektrotechnik, Berlin*, Nov. 1948, Vol. 2, No. 11, pp. 297–304.) A detailed theory is given of the directional characteristics of arrays of vertical aerials over uneven ground. A d.f. system described by Stenzel, while giving good results for eliminating the effects of variations of earth constants in the case of level ground, does not give such good results over uneven ground. Measurements of directional characteristics which support the theory are described and the physical connection between measured reflection factors and the electrical properties of the ground is considered.

621.396.677.029.54 **40**

Design Considerations for Directive Antennae-Arrays at Medium-Wave Broadcast Frequencies, taking into account the Final Radio-Frequency Amplifier Circuits.—J. C. Nonnekens. (*H.F., Brussels*, 1949, No. 3, p. 80. In English.) Correction to 1605 of 1949.

CIRCUITS AND CIRCUIT ELEMENTS

061.4 : [621.317.7 + 621.38 + 621.396.69] **41**

Radiolympia Review.—(*See* 145.)

621.3 : 512.974 **42**

Vectorial Space. Region of Representation for Electrotechnics.—Nasse. (*See* 132.)

621.314.3† **43**

The Direct-Current Choke.—W. Blankenburg. (*Elektrotechnik, Berlin*, May 1949, Vol. 3, No. 5, pp. 135–140.) Starting from the ordinary premagnetized choke, magnetic amplifier arrangements are developed and their properties are investigated. A null-current model is described with an amplification factor of about 10^7; this is controlled by a thermoelement and gives a load power of about 1 W.

621.314.3† **44**

An Experimental Study of the Magnetic Amplifier and the Effects of Supply Frequency on Performance.—E. H. Frost Smith. (*J. Brit. Instn Radio Engrs*, Oct. 1949, Vol. 9, No. 10, pp. 358–373.) Magnetic amplifiers have hitherto been used mainly for amplifying small d.c. powers. The a.c. supply to the amplifier then operates normally at power frequencies. The time constant of the amplifier is limited by the supply frequency; the possibility of reducing response time by using a supply frequency of about 20 kc/s is considered. For a given frequency and power output, there is an optimum core size, which becomes more critical as the supply frequency is increased. Experimental results show that the magnetic amplifier has distinct possibilities as an a.f. amplifier.

621.314.3† **45**

Magnetic Amplifiers.—(*Proc. Instn elect. Engrs*, Part II, Oct. 1949, Vol. 96, No. 53, pp. 767–768.) Report of an I.E.E. Measurements Section discussion on 2728 of 1949 (Milnes) and 2729 of 1949 (Gale & Atkinson).

621.314.3† **46**

The Magnetic Amplifier.—P. M. Kintner & G. H. Fett. (*Radio & Televis. News, Radio-Electronic Engng Supplement*, Aug. 1949, Vol. 13, No. 2, pp. 14–18, 28.) An analysis from first principles, with discussion of factors affecting design.

621.314.3† **47**

Magnetic Amplifiers.—M. Alixant. (*Radio tech. Dig., Edn franç.*, June 1949, Vol. 3, No. 3, pp. 153–159. Bibliography, pp. 159–163.) Short account of basic principles, practical construction and applications.

621.316.86 **48**

Electrolytic Thermistors.—F. Gutmann & L. M. Simmons. (*Rev. sci. Instrum.*, Sept. 1949, Vol. 20, No. 9, pp. 674–675.) A thermistor for use with a.c. is easily constructed by immersing Pt electrodes in a viscous solution of waterglass. A temperature change of only 7.2° at 300° K will double or halve the resistance. This thermistor is three times as sensitive as commercial thermistors, withstands alternating potentials of more than 240 V, and its resistance and sensitivity when cold are easily controlled. Other solutions were tested; they could withstand high alternating potentials, but were less sensitive.

621.316.86.001.8 **49**

Properties and Applications of Thermistors.—E. Ancel. (*Radio franç.*, Sept. 1949, No. 9, pp. 3–10.) Applications to thermometry, temperature compensation and control, power measurement, voltage regulation, a.g.c., and in relays, oscillators and modulators, are outlined.

621.318.2 : 621-23 50

Permanent Magnets in Drag Devices and Torque-Transmitting Couplings.—R. J. Parker. (*Gen. elect. Rev.*, Sept. 1949, Vol. 52, No. 9, pp. 16–20.) Permanent magnets are used in four principal types of torque-transmitting device, namely: (*a*) eddy-current current devices, (*b*) hysteresis devices, (*c*) iron/oil magnetic couplings, (*d*) salient-pole synchronous couplings. Specimens of each of these types are briefly described, and formulae for the torque are obtained.

621.318.42 : 621.314.263 : 621.314.3† 51

The Use of Ferrite-Cored Coils as Converters, Amplifiers and Oscillators.—V. D. Landon. (*RCA Rev.*, Sept. 1949, Vol. 10, No. 3, pp. 387–396.) Theoretical treatment of the behaviour of nonlinear inductors used as frequency converters up to a few Mc/s shows that (*a*) if the excitation frequency is higher than the signal frequency and the i.f., the circuit regenerates; (*b*) if the excitation is sufficient, the circuit may oscillate at the signal frequency and i.f. simultaneously; (*c*) if either the r.f. or the i.f. circuit is tuned to a frequency above the oscillator frequency, the circuit is degenerative. Experimental results and transformer specifications are given, and possible applications are noted.

621.318.423.011.3 52

Inductance of Air-Cored Single-Layer Cylindrical Coils.—K. Schönbacher. (*Elektrotechnik, Berlin*, Oct. 1949, Vol. 3, No. 10, pp. 327–329.) A simple formula involving the ratio of the coil diameter d to its length l is derived from expressions given by Rayleigh and by Lorenz. For all practical cases the use of correction formulae can be avoided by replacing l by $l + 0.45d$.

621.318.572 : 621.396.645 : 537.311.33 : 621.315.59 53

A Transistor Trigger Circuit.—H. J. Reich & K. L. Ungvary. (*Rev. sci. Instrum.*, Aug. 1949, Vol. 20, No. 8, pp. 586–588.) A trigger circuit may be formed by inserting a suitable resistor in the lead connected to the base of a transistor. Such circuits are in many ways superior to valve circuits; a circuit with a triggering time of 0.1 μs giving an output of 5–6 V was found to be stable at frequencies up to 1 Mc/s; stability at frequencies up to at least 10 Mc/s is probable. The circuit may be converted into a relaxation oscillator or a pulse generator.

621.392 54

A Network Bisection Theorem.—V. D. Landon. (*RCA Rev.*, Sept. 1949, Vol. 10, No. 3, pp. 448–450.) It is shown how any given symmetrical ladder network may be split into two simpler networks such that the transfer factor of the whole is equal to half the product of the transfer factors of the parts, as here defined.

621.392 : 621.3.015.3 55

The Effect of Pole and Zero Locations on the Transient Response of Linear Dynamic Systems.—J. H. Mulligan, Jr. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, p. 1181.) Corrections to 2165 of 1949.

621.392.5 : 621.392.26† 56

Basis of the Application of Network Equations to Waveguide Problems.—D. M. Kerns. (*Bur. Stand. J. Res.*, May 1949, Vol. 42, No. 5, pp. 515–540.) A mathematical and fundamental paper. The solution of network problems can be effected by means of a matrix equation in which the constants are parameters characteristic of the components, and the variables are currents and voltages. The physical processes inside the components are not explicitly considered. A matrix equation is developed to enable this technique to be extended to waveguide and transmission-line problems; here tangential electric and magnetic fields replace the currents and voltages as variables. Many of the results obtained are identical with those of the theory of quadripoles. In particular, a reciprocity theorem and Foster's reactance theorem are considered.

621.392.52 57

Error-Actuated Power Filters.—G. Newstead & D. L. H. Gibbings. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1115–1119.) Voltages derived from the harmonic voltages of an a.c. source of impure waveform are applied through a negative-feedback path to cancel unwanted harmonics. The method could be extended to apply to a given band of frequencies. Large powers can thus be controlled by means of components of much lower rating. Design equations are given and the operation of a typical filter circuit is examined.

621.392.52 58

Smoothing Circuits : Part 2 — Inductance-Capacitance.—"Cathode Ray". (*Wireless World*, Nov. 1949, Vol. 55, No. 11, pp. 418–422.) Continuation of 3371 of 1949. The calculation of hum voltages is discussed.

621.392.52 59

Mechanical Filters for Radio Frequencies.—W. van B. Roberts & L. L. Burns, Jr. (*RCA Rev.*, Sept. 1949, Vol. 10, No. 3, pp. 348–365.) The theory of neck-type and slug-type filters is discussed. A simple type of band-pass filter composed of loosely-coupled metal resonators with magnetostrictive drive and take-off has very sharp frequency discrimination and is readily constructed for frequencies up to $\frac{1}{2}$ Mc/s and bandwidth less than 3% of the mid-frequency. The voltage gain of an amplifier stage using such a filter is generally considerably lower than that of a stage using electrical circuit coupling.

621.392.52 : 621.3.015.3 60

Transient Response of Filters.—M. S. Corrington. (*RCA Rev.*, Sept. 1949, Vol. 10, No. 3, pp. 397–429.) The ideal low-pass filter is defined as one whose amplitude response is constant at all frequencies between zero and cut-off, and falls off at a fixed rate beyond cut-off. The flat characteristic of such a filter is not necessarily desirable when the rise time and the amount of overshoot are important. The transient response of an ideal filter has been computed, and the results applied to the determination of the transient response for systems with selectivity curves composed of straight-line segments. Relations between the transient response of a low-pass filter and that of the corresponding high-pass filter are considered.

621.392.52.012.8 61

Generalized Theory of the Band-Pass Low-Pass Analogy.—P. R. Aigrain, B. R. Teare, Jr. & E. M. Williams. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1152–1155.) Laplace transforms are used. Video-circuit equivalents are shown to exist for asymmetrical systems with low-level modulation as well as for symmetrical systems. Several illustrations are given.

621.392.6 62

Synthesis of Passive Networks with Any Number of Pairs of Terminals, given their Impedance or Admittance Matrices.—M. Bayard. (*Bull. Soc. franç. Elect.*, Sept. 1949, Vol. 9, No. 96, pp. 497–502.) The necessary and sufficient condition is established for a square symmetric matrix to be an impedance or an admittance matrix. The condition is that the quadratic form associated with the given matrix is a "positive real function" in the sense in which Brune uses the term. This result generalizes the theorem established by Brune for the impedance matrices of 2-pole networks (1932 Abstracts, p. 280) and by Gewertz for those

of quadripoles (1934 Abstracts, p. 514). The greater part of the mathematical treatment included in the original paper presented at a meeting of the Société française des Électriciens is here omitted.

621.396.611.1 **63**

Energy Fluctuations in a van der Pol Oscillator.—N. Minorsky. (*J. Franklin Inst.*, Sept. 1949, Vol. 248, No. 3, pp. 205–223.) The behaviour of the van der Pol oscillator, whose differential equation is

$$\ddot{x} - \epsilon(1 - x^2)\dot{x} + x = 0,$$

resembles that of the harmonic oscillator when ϵ is small; the very different behaviour when ϵ is large is examined with the aid of data obtained by the method of isoclines.

The van der Pol oscillator has a prescribed energy content which does not depend on the initial conditions, whereas the harmonic oscillator can have any energy content. *The van der Pol oscillator is* nevertheless the best means available for physical realization of simple harmonic oscillation if ϵ is small. The perturbation method is used to determine the fluctuations of energy with time and azimuth for $\epsilon \ll 1$. For $\epsilon \gg 1$, the energy fluctuations of the van der Pol oscillator are quasi-discontinuous, and somewhat analogous to those of a pneumatic hammer. See also 1930 Abstracts, p. 503 (van der Pol).

621.396.611.1 : 621.3.015.33 **64**

Response of RC Circuits to Multiple Pulses.—D. Levine. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1207–1208.) The voltage across the capacitor due to the application of several pulses in succession is determined by a generalized step-by-step process for any number of pulses. Only the well-known charge/discharge functions for a d.c. circuit are required. The voltage across the resistor can then be directly determined.

621.396.611.1.029.63 **65**

Tank Circuits as Resonators in Decimetre-Wave Technique.—G. Megla. (*Elektrotechnik, Berlin*, Nov. 1948, Vol. 2, No. 11, pp. 305–312.) Detailed treatment of the properties of tank circuits, with illustrations of many different types and derivation of suitable formulae for the resonance frequency and resistance. Below a wavelength of about 10 cm the Q of tank circuits decreases, so that cavity resonators are preferable.

621.396.615 : 621-12 **66**

The Reciprocator.—W. C. White & H. W. Lord. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 70–71.) A ring oscillator, comprising 2 one-shot multivibrators whose on and off periods can be varied independently between 0.2 and 1.5 sec, energizes linear or rotary solenoids which generate reciprocating motion in which the distance/time relationship during the whole stroke can be controlled.

621.396.615.141.1 : 2 **67**

On the Theory of Magnetron Barkhausen Oscillations.—H. G. Möller. (*Elektrotechnik, Berlin*, May 1949, Vol. 3, No. 5, pp. 129–133.) The mechanism of the production of Barkhausen oscillations in whole-anode magnetrons is similar to that for triodes. For magnetrons the period of the pendular electrons is shorter than that of the retarded electrons, the reverse being the case for triodes. In consequence of the different frequencies, phase focusing occurs. An explanation of the oscillation mechanism is put into a mathematical form and the approximate path of the electrons, the dependence of the pendular frequency on the amplitude, the a.c. contribution to the space charge, and the excitation factor, are calculated. An upper limit for the efficiency can be found if it is assumed that the total pendular energy can be converted to oscillation energy.

621.396.615.17 : 621.317.755 **68**

Panoramic Sweep Circuits.—C. B. Clark & F. J. Kamphoefner. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 111–114.) Brief details, with block diagrams, of 12 methods of obtaining sweep voltages for panoramic receivers, f.m. signal generators and r.f. spectrum analysers.

621.396.619.23 : 621.396.615.17 **69**

A Modulator Producing Pulses of 10^{-7} Second Duration at a 1-Mc/s Recurrence Frequency.—M. G. Morgan. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, p. 1178.) Correction to 2185 of 1949.

621.396.645 **70**

Musician's Amplifier.—D. Sarser & M. C. Sprinkle. (*Audio Engng, N. Y.*, Nov. 1949, Vol. 33, No. 11, pp. 11–13, 55.) An adaptation of the Williamson circuit (2715 of 1947 and 3101 of 1949). Williamson's output-transformer performance specifications are satisfactorily met by the Peerless transformer Type S-265Q. Construction and performance details are discussed and shown graphically. Quality is outstanding.

621.396.645 **71**

Amplification by Direct Electronic Interaction in Valves without Circuits.—P. Guénard, R. Berterottière & O. Doehler. (*Bull. Soc. franç. Élect.*, Oct. 1949, Vol. 9, No. 97, pp. 543–549.) See 2977 of 1949.

621.396.645 **72**

Simplified Preamplifier Design.—H. T. Sterling. (*Audio Engng, N. Y.*, Nov. 1949, Vol. 33, No. 11, pp. 16–17, 45.) Unnecessary complications may be avoided by making practical compromises in performance requirements. A modified Pickering circuit is recommended. See also 1608 of 1948 (Burwen).

621.396.645 : 539.17 **73**

Fast Pulse-Amplifiers for Nuclear Research.—W. C. Elmore. (*Nucleonics*, Sept. 1949, Vol. 5, No. 3, pp. 48–55.) Discussion of the limitations affecting the speed of response of amplifiers without feedback, and of the design of amplifiers whose speed of response approaches the theoretical limit. In each stage of the amplifier, the output signal is assumed to rise monotonically to its final steady value for a step-voltage input. The relations between the overall rise time of an amplifier and the rise times of individual stages are considered. Characteristics of 11 valves useful for pulse amplifiers are tabulated and discussed.

621.396.645 : 578.088.7 **74**

The Design and Construction of an Amplifier for Bio-Electric Recording.—R. T. Jamieson. (*Trans. S. Afr. Inst. elect. Engrs*, Sept. 1949, Vol. 40, Part 9, pp. 204–212. Discussion, p. 212.) Requirements of such amplifiers are discussed. An a.c. mains-operated amplifier is described. The pre-amplifier has a gain of 74 db and consists of two coupled-cathode stages followed by a cathode follower with gas-valve coupling to the input of the high-level amplifier, which is a 3-stage resistance-coupled amplifier with cathode-follower output and a gain of 66 db. Full-scale deflection on the c.r.o. is obtained for a 10-μV input signal. The two amplifiers should be isolated in places reasonably free from electrical noise.

621.396.645 : 621.385 **75**

Operation of Output Valves in High-Power Public-Address Amplifiers.—N. L. Bezladnov. (*Radiotekhnika, Moscow*, July/Aug. 1949, Vol. 4, No. 4, pp. 8–17. In Russian.) Continuation of 2602 of 1949. The efficiency of the final stage of such amplifiers can be improved by (*a*) anode loading, (*b*) use of a limiting device. In case (*a*), the average anode dissipation under dynamic operating conditions, not the maximum dissipation, should be regarded as determining maximum permissible

anode loading. In case (*b*), the transmission power can be increased without increasing nonlinear distortion. The theory of both methods is discussed and experimental curves are shown.

621.396.645.37 : 518.4 **76**

Calculator and Chart for Feedback Problems.—J. H. Felker. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1204–1206.)

621.396.645.37.001.4 : 621.3.016.352 : 621.3.015.3 **77**

Examination of Amplifier Stability by applying a Sudden D.C. Test Voltage.—B. Carniol. (*Tesla tech. Rep., Prague*, March 1949, pp. 11–20.) Nyquist's theorem and the corresponding characteristic curve for feedback amplifiers are discussed and practical methods for determining the characteristic for any particular amplifier are described in detail. The computation accuracy depends on the exactness with which the phase characteristic can be represented by a series of straight lines. The advantages of the method of testing amplifier stability by studying the response to a transient voltage or square wave are considered. The effect of circuit variations in enhancing or reducing transient oscillations in the output of various amplifiers is illustrated by a series of oscillograms. A Nyquist characteristic for one of the amplifiers shows the existence of a critical frequency near which a definite tendency to instability exists. See also 218 below.

621.396.645.371 : 621.3.015.3 **78**

Transient Phenomena in Wide-Band Feedback Amplifiers.—O. B. Lur'e. (*Zh. tekh. Fiz.*, Aug. 1949, Vol. 19, No. 8, pp. 952–972. In Russian.) Negative-feedback amplifiers should be widely used in video circuits although they give somewhat lower gain than that obtained with the usual distortion-correcting circuits. Formulae are derived for determining circuit parameters when given (*a*) the time for reaching the steady state, (*b*) the maximum permissible overshoot.

621.396.662 : 621.396.61 **79**

A New Type of V.H.F. Tank Design.—B. E. Parker. (*FM-TV*, Oct. 1949, Vol. 9, No. 10, pp. 14–15.) A balanced 2-wire section of transmission line is tuned by varying its distance from a flat conductor. This affects the series inductance and the shunt capacitance, and hence the surge impedance. This arrangement has advantages in a push-pull output circuit, including the fact that the tuning element is at earth potential.

621.396.662 : 621.396.61 : 513.76 **80**

Bilinear Transformations applied to the Tuning of the Output Network of a Transmitter.—K. S. Kunz. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1211–1217.) The theory of bilinear transformations in a complex plane, by which circles are transformed into circles, is used to determine the actual value C of the secondary capacitance for (*a*) resonance of the secondary circuit, (*b*) a maximum or minimum of the average anode current, (*c*) maximum coupled-circuit efficiency. (*b*) and (*c*) require values of C differing widely from that required for (*a*), except for the case of a series load in the secondary circuit. Although the values of C required for (*b*) and (*c*) may differ considerably, the difference is small if the *Q* of the primary coil is large and the coupling coefficient very small.

621.396.662.2 **81**

A New Decade Inductor.—H. W. Lamson. (*Gen. Radio Exp.*, July 1949, Vol. 24, No. 2, pp. 1–8.) For the range 1 mH–10 H, intended for use at audio and the lower ultrasonic frequencies. Each decade unit consists of four inductors whose inductances are in the ratio 1 : 2 : 2 : 5, with an overall e.s. shield and a special switch for connecting the inductors in series as required. Multi-layer uniformly wound toroidal coils are used. These have a banked winding to minimize distributed capacitance and can be stacked in a compact unit with negligible mutual inductance. The stabilized Mo-permalloy dust cores have low eddy-current and hysteresis losses. Construction details and electrical properties are discussed.

GENERAL PHYSICS

530.12 : 531.18 **82**

The Special Relativity Theory.—E. Kübler. (*Elektrotechnik, Berlin*, Nov. 1948, Vol. 2, No. 11, pp. 323–327.) A simple explanation, by an electrician for electricians, of the electrodynamic consequences of this theory.

534.26 + 535.42. **83**

On the Theory of Diffraction by an Aperture in an Infinite Plane Screen : Part 1.—H. Levine & J. Schwinger. (*Phys. Rev.*, 15th Oct. 1948, Vol. 74, No. 8, pp. 958–974.) The case of a scalar plane wave is considered. The wave function at an arbitrary point in space is expressed in terms of its values in the aperture, and constructed so as to vanish on the screen, in accordance with the assumed boundary condition. An integral equation is obtained, by means of which the amplitude of the diffracted spherical wave at large distances from the aperture can be expressed in a form which is stationary with respect to small variations of the aperture fields arising from a pair of incident waves. This expression is independent of the scale of the aperture fields. The transmission cross-section of the aperture for a plane wave is simply related to the diffracted amplitude observed in the direction of incidence. The method is applied to the case of a wave incident normally on a circular aperture, for which exact results are available; by using suitable trial aperture fields in the variational method, approximate but sufficiently accurate results are obtained for the diffracted amplitude and transmission cross-section over a wide range of frequencies. See also 3294 of 1948 (Spence).

534.26 + 535.42 **84**

On the Transmission Coefficient of a Circular Aperture.—C. J. Bouwkamp : H. Levine & J. Schwinger. (*Phys. Rev.*, 15th May 1949, Vol. 75, No. 10, pp. 1608–1609.) Comment on 83 above and the authors' reply.

535.42 : 538.566 **85**

New Solution of the Problem of the Diffraction of Electromagnetic Waves by a Plane Perfectly Conducting Screen.—J. P. Vasseur. (*C. R. Acad. Sci., Paris*, 19th Sept. 1949, Vol. 229, No. 12, pp. 586–587.) Using the same notation as in a previous note (3124 of 1949), a rigorous expression of Babinet's principle is derived by the help of integral equations, the integration extending over the surface of the screen. This expression of Babinet's principle has been given by H. G. Booker in an unpublished paper. See also 1335 of 1947 (Booker).

537.311.62 **86**

Skin Effect.—C. Vazaca. (*Radio tech. Dig., Edn franç.*, June–Oct. 1949, Vol. 3, Nos. 3–5, pp. 175–184, 223–233 & 307–316.) Reprint of article in *Bull. sci. Recherches technol. Roumanie*, 1948, Vol. 3, Nos. 3/4. The concentration of the current in the surface layers of a conductor carrying a.c. is quite analogous to the concentration of magnetic field in the surface layers under the action of a longitudinal alternating magnetic field, owing to eddy-current effects. Starting from Maxwell's equations, formulae are developed for the case of a long cylindrical conductor, giving the law of distribution within the conductor of the current density and total current, and also of the magnetic field and total magnetic flux. Formulae are also derived for

the impedance and the complex reluctance, the effective depth of penetration of the current and of the magnetic field, and the energy dissipated per unit length or volume of the conductor.

537.521.7 87

The Positive Streamer Mechanism of Spark Breakdown.—W. Hopwood. (*Proc. phys. Soc.*, 1st Oct. 1949, Vol. 62, No. 358B, pp. 657–664.)

537.525 : 538.551.25 88

Theory of Plasma Oscillations. A. Origin of Medium-Like Behavior.—D. Bohm & E. P. Gross. (*Phys. Rev.*, 15th June 1949, Vol. 75, No. 12, pp. 1851–1864.) A theory of electron oscillations of an unbounded plasma of uniform ion density is developed, taking into account the effects of random thermal motions, but neglecting collisions. The frequencies are determined at which a plasma can undergo organized steady-state small oscillations to which a linear approximation applies. Organized oscillations of wavelength smaller than the Debye length for the electron gas are not possible. For a given wavelength, a plasma can oscillate with arbitrary frequency, but the frequencies not satisfying a 'steady-state dispersion relation' describe motions in which, after some time, there is no contribution to large-scale averages. The treatment is extended to the case of large steady-state oscillations for which the equations are nonlinear, and the general nature of solutions for this case is discussed. See also 89 below.

537.525 : 538.551.25 : 621.396.822 89

Theory of Plasma Oscillations. B. Excitation and Damping of Oscillations.—D. Bohm & E. P. Gross. (*Phys. Rev.*, 15th June 1949, Vol. 75, No. 12, pp. 1864–1876.) The theory discussed in 88 above is extended to include the effects of collisions and of special groups of particles having well-defined ranges of velocities. A wave tends to be damped by collisions in a time of the order of the mean time between collisions. Groups of particles having speeds far above the mean thermal speed may cause instability only limited by effects neglected in the linear approximation. In the absence of plasma oscillations, any beam of well-defined velocity is scattered by the individual plasma electrons acting at random, but the scattering is much greater when all the particles oscillate in unison. The beams are thus scattered by the oscillations they produce. This type of instability may be responsible for solar and cosmic r.f. noise.

538.12 90

Production of Intense Magnetic Fields by means of Pulses.—G. Raoult. (*Ann. Phys., Paris*, July/Aug. 1949, Vol. 4, pp. 369–421.) By means of recurrent voltage pulses, a suitable commutator and a grid-controlled Hg rectifier, a capacitor is charged and then discharged through a coil about 100 times per sec. In this way magnetic field pulses of roughly sinusoidal shape and of duration about 15 μs have been produced with a peak power of the order of 90 MW and a mean power of 850 W. The maximum field strength was 52 000 gauss in a useful volume of about 3 cm^3. Applications to phenomena of polarization rotation and magnetic double refraction in CS_2 are described.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523 : 621.396.9 91

Radio Astronomy.—J. S. Hey. (*Mon. Not. R. astr. Soc.*, 1949, Vol. 109, No. 2, pp. 179–211. Bibliography, pp. 211–214.) General survey of (*a*) radio detection of meteors and reflections from the moon, (*b*) solar radiation, and (*c*) galactic radiation.

523.746.5 92

The Probable Behavior of the Next Sunspot Cycle.—W. Gleissberg. (*Astrophys. J.*, July 1949, Vol. 110, No. 1, pp. 90–92.) The probability laws of sunspot variations, which have yielded successful predictions for the present sunspot cycle, suggest that a steep ascent to a very high maximum is highly probable in the next cycle, and that the period of low activity preceding the next cycle will be extremely short.

550.385"1949.01.24/.26" 93

Recurrence Features of the Magnetic Storm of 1949 January 24–26.—(*Observatory*, Oct. 1949, Vol. 69, No. 852, pp. 195–196.) Great magnetic storms, beginning with a 'sudden commencement' and associated with solar flares, do not usually show any recurrence connected with the solar rotation. This particular storm, however, recurred with sudden commencements on 21st Feb. 1949 and 21st March 1949. No positive evidence of flares associated with the last two storms is available. These storms were of the polar type, and were associated with aurorae. Magnetograms are reproduced and compared.

551.51 94

The Earth's Upper Atmosphere.—D. R. Bates. (*Mon. Not. R. astr. Soc.*, 1949, Vol. 109, No. 2, pp. 215–242. Bibliography, pp. 242–245.) Known facts about the constitution, density and temperature at various levels are summarized, and methods of exploring the upper atmosphere are surveyed. The ionized layers and aurorae are also discussed.

551.510.535 95

Ionization in the Earth's Upper Atmosphere.—(*Observatory*, Oct. 1949, Vol. 69, No. 852, pp. 185–191.) Report of a discussion held at Manchester University on 2nd July 1949.

551.557 : 621.396.9 96

A Radio Method of Measuring Winds in the Ionosphere.—S. N. Mitra. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 441–446.) A pulse transmitter operating at a frequency of about 4 Mc/s is used in conjunction with 3 receivers at the corners of a right-angled triangle with sides of the order of 100 m. A single magneto-ionic component received at one point will show fading only if the ionosphere changes either by random motion of irregularities or by steady motion of the ionosphere as a whole. A uniform drift of the ionosphere would cause the diffraction pattern to move past a receiver with twice the velocity of the drift. Wind velocities can thus be deduced when the records at the 3 receivers are similar but displaced in time. Results thus obtained are compared with those of previous measurements by other methods.

551.594.6 97

Large-Scale Variation of the Level of Atmospherics during the Antarctic Cruise of the *Commandant Charcot*.—R. Bureau & M. Barré. (*C. R. Acad. Sci., Paris*, 26th Sept. 1949, Vol. 229, No. 13, pp. 626–627.) The recorder used on board the *Commandant Charcot* has been calibrated by Carbenay's method (2902 of 1948). A map is given showing, for the various sections of the whole course, the operational threshold in maxwells per metre. Longitude and latitude effects are clearly indicated and the difference between the threshold sensitivities for January and March in Australian waters indicates also a seasonal effect. See also 3429 of 1949.

551.594 98

Atmospheric Electricity. [Book Review]—J. A. Chalmers. Publishers: Oxford University Press, Oxford, 1949. 175 pp., 15s. (*Proc. phys. Soc.*, 1st Oct. 1949, Vol. 62, No. 358B, pp. 665–666.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 99

Radar.—(*Bull. Soc. franç. Elect.*, Oct. 1949, Vol. 9, No. 97, p. 532.) In Quillet's encyclopaedia the following entries may be found :—

Radar.—En Perse : Membre d'une milice créée pour protéger les voyageurs.

[*In Persia* : Member of a militia established for the protection of travellers.]

Radarie.—Droit persan payé par les marchandises sous la protection des radars.

[Persian toll paid on merchandise under the protection of radars.]

621.396.9 : 061.4 100

Radar Research and Development Establishment [Malvern].—(*Engineer, Lond.*, 30th Sept. 1949, Vol. 188, No. 4888, p. 373.) A brief survey of work in progress at this establishment, based on a visit made on a recent open day. See also *Nature, Lond.*, 29th Oct. 1949, Vol. 164, No. 4174, pp. 740–741.

621.396.9 : 621.385.832 : 535.371.07 101

Radar Screens.—de la Pinsonie. (*See* 267.)

621.396.933 : 526.92 102

Distance-Measuring Equipment for Aircraft Navigation.—V. D. Burgmann. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 395–402.) An airborne light-weight radar set works in conjunction with a responder beacon on the ground. The radar transmitter sends out a stream of pulses which are returned by the beacon at a slightly different frequency. The time interval between the outgoing and corresponding return pulse is measured automatically and shown on a meter calibrated in miles. The design of the airborne set and results obtained in trials on an air route are discussed. Well-established techniques are used.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.37 103

Review of the Interpretations of Luminescence Phenomena.—F. E. Williams. (*J. opt. Soc. Amer.*, Aug. 1949, Vol. 39, No. 8, pp. 648–654.)

535.37 104

A Survey of Present Methods used to Determine the Optical Properties of Phosphors.—W. B. Nottingham. (*J. opt. Soc. Amer.*, Aug. 1949, Vol. 39, No. 8, pp. 641–647.)

535.37 105

The Intensity Dependence of the Efficiency of Fluorescence of Willemite Phosphors.—W. Hoogenstraaten & F. A. Kröger. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 541–556. In English.)

535.37 106

Temperature Quenching and Decay of Fluorescence in Zinc and Zinc-Beryllium Silicates activated with Manganese.—F. A. Kröger & W. Hoogenstraaten. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 557–568. In English.) Continuation of 3156 of 1949 and 105 above.

535.37 : 546.412.84 107

Optical Properties of Calcium Silicate Phosphors.—F. J. Studer & G. R. Fonda. (*J. opt. Soc. Amer.*, Aug. 1949, Vol. 39, No. 8, pp. 655–660.)

535.37 : 546.471.61 108

A Correlation between Cathodoluminescence Efficiency and Decay [of Mn-activated ZnF_2] **as a Function of Temperature.**—R. H. Bube. (*J. opt. Soc. Amer.*, Aug. 1949, Vol. 39, No. 8, pp. 681–684.)

535.37 : 546.472.21 109

Sodium and Lithium as Activators of Fluorescence in Zinc Sulfide.—F. A. Kröger. (*J. opt. Soc. Amer.*, Aug. 1949, Vol. 39, No. 8, pp. 670–672.)

535.37 : 546.472.21 110

The Luminescence of Zinc Sulfide Activated by Lead.—N. W. Smit & F. A. Kröger. (*J. opt. Soc. Amer.*, Aug. 1949, Vol. 39, No. 8, pp. 661–663.)

537.311.31 + 536.212.2] : 546.78 111

Thermal and Electrical Resistance of a Tungsten Single Crystal at Low Temperatures and in High Magnetic Fields.—J. de Nobel. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 532–540. In English.)

537.312.62 : 621.396.622 112

Detection at Radio Frequencies by Superconductivity.—Lebacqz, Clark, Williams & Andrews. (*See* 223.)

538.221 113

Magnetic Materials for Electrical Power Plant.—F. Brailsford. (*Engineering, Lond.*, 16th Sept. 1949, Vol. 168, No. 4364, pp. 293–296.) Long summary of paper read before the British Association. The progress in steel- and sheet-making processes is reviewed. Different steels are compared; the need for improving magnetic quality is stressed. Methods outlined include directional cold-rolling technique and a laboratory method for preferred orientation in silicon-iron crystals. For other summaries see 3442 of 1949 and *Electrician*, 2nd Sept. 1949, Vol. 143, No. 3716, pp. 737–738. For discussion, see *Electrician*, 16th Sept. 1949, Vol. 143, No. 3718, pp. 893–894.

549.514.51 114

The Quartz Crystal as Raw Material for High-Frequency Technics.—H. Neels. (*Elektrotechnik, Berlin*, Oct. 1949, Vol. 3, No. 10, pp. 318–322.) Discussion of the various forms of natural quartz, methods of selection by visual inspection, and optical and electrical tests for twinning.

549.514.51 115

A New Crystal Cut for Quartz with Zero Temperature Coefficient.—E. J. Post. (*Tijdschr. ned. Radiogenoot.*, Sept. 1949, Vol. 14, No. 5, pp. 147–157.) A rotated y-cut making an angle of about $+36°$ with the z-axis. It is suitable for the frequency range 400–1 000 kc/s and the term HT-cut is proposed. Resonators can be mounted by soldering springs at the nodal points. Spurious responses are absent in the range within 10% above or below the main resonance frequency.

549.514.51 : 621.392.52 116

Crystals for Electrical Filters.—R. Taylor, R. Bechmann & A. C. Lynch. (*Research, Lond.*, Sept. 1949, Vol. 2, No. 9, pp. 414–417.) The equivalent circuit of a quartz crystal is discussed, and factors that must govern the choice of suitable crystals for practical applications are considered. A crystal may have as many as 18 piezoelectric constants, 6 permittivities and 21 elastic constants, all of which may have different temperature coefficients. The highest piezoelectric constants for various classes of piezoelectric crystal are tabulated. Temperature coefficients, ferroelectric properties, methods of growing crystals, etc., are briefly discussed. The ideal piezoelectric material has yet to be found.

621.315.59 + 621.314.6 117

Semiconductor Rectifiers.—(*See* 240.)

621.315.59 + 621.314.63 118

Semi-Conductors and Rectifiers.—N. F. Mott. (*Proc. Instn elect. Engrs*, Part I, Sept. 1949, Vol. 96, No. 101, pp. 253–260.) Full paper : summary noted in 2814 of 1949.

621.315.59 : 621.315.61 119

New Dielectric Materials : Semi-Conductors, Plastics and Silicones.—(*Electrician*, 2nd Sept. 1949, Vol. 143, No. 3716, pp. 735–736. Discussion, *ibid.*, 16th Sept. 1949, Vol. 143, No. 3718, pp. 803–804.) Summary of paper presented before the British Association by R. W. Sillars, entitled "Insulating and Semi-Conducting Materials". Another summary noted in 3442 of 1949. See also 3168 of 1949 (Jones, Scott & Sillars).

621.315.612 120

Titanates with High Dielectric Constant.—H. Sachse. (*Z. angew. Phys.*, Aug. 1949, Vol. 1, No. 10, pp. 473–484.) A comprehensive review of the results of measurements, made in various countries, of the properties of a wide range of titanate materials. 57 references are given.

621.315.612 : 621.317.37.029.6 121

The Measurement of the Dielectric Properties of High-Permittivity Materials at Centimetre Wavelengths.—J. G. Powles & W. Jackson. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 383–389.) Results of room-temperature measurements on ceramic specimens of the titanates of Mg, Ca, Sr and Ba at 1.5 Mc/s and at 9 450 Mc/s, and on $BaTiO_3$ at 24 000 Mc/s are discussed. $BaTiO_3$ is unique among these materials, because a considerable fall in permittivity and a large increase in $\tan\delta$ occur at the higher frequencies. Variable-temperature measurements on $BaTiO_3$ show that the crystallographic change associated with the Curie point at 120 °C affects the permittivity at 9 450 Mc/s and at 1.5 Mc/s in an analogous manner. Measurements at the same frequencies on (Ba,Sr) titanate compositions are also described: the composition 56% $BaTiO_3$: 44% $SrTiO_3$ has a permittivity of 760 and $\tan\delta = 0.02$ at 9 450 Mc/s and 20 °C. The three methods of measuring high permittivities at cm λ developed in the course of this work are described.

621.315.614 : 621.317.372 122

Electrical Testing of Capacitor Paper.—Endicott. (*See* 144.)

621.315.614.62/.63 : 621.319.4 123

Metallizing Paper for Capacitors.—H. G. Wehe. (*Bell Lab. Rec.*, Sept. 1949, Vol. 27, No. 9, pp. 317–321.) A German wartime metal-vapour spraying process, developed by the Bell Telephone Laboratories. For a given capacitance the metallized-paper capacitor is much smaller than the corresponding metal-foil capacitor, and has the advantage of self-healing of punctures through the insulation.

621.315.615.011.5 : 537.226.2 124

A Contribution to the Theory of the Dielectric Constant of Polar Liquids.—T. G. Scholte. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 437–449. In English.)

621.315.615.2.011.5 : 537.226.2 125

The Absolute Dielectric Constant of Benzene.—F. van der Maesen. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 481–483.) The dielectric constant is appreciably reduced by prolonged drying over Na. This confirms the conclusion of Hartshorn & Oliver (*Proc. roy. Soc. A*, 1929, Vol. 123, pp. 664–685) that moisture is the only impurity of importance in dielectric-constant measurements. Values obtained differ by less than 0.1% from the generally accepted standard values. The method of purification is discussed.

621.315.616 126

Ethoxylines.—E. Preiswerk & C. Meyerhans. (*Elect. Mfg, N.Y.*, July 1949, Vol. 44, No. 1, pp. 78–81, 166.) Properties and applications are discussed. The first resins of this type were supplied under the trade name Araldite; for another account see 3450 of 1949 (Moss).

621.315.616 : 546.287 127

Recent Progress made in Silicone Rubber Materials.—C. E. Arntzen & R. D. Rowley. (*Materials & Methods*, Oct. 1949, Vol. 30, No. 4, pp. 73–76.) The chemical nature of silicones, mechanical and physical properties of various silastic rubbers, and applications, are discussed. Silicone rubbers are much less affected by temperatures of the order of 300 °F than natural rubbers.

621.315.618.011.5 : 537.226.2 128

The Absolute Dielectric Constant of Gases at Pressures of 0–80 Atm. at 25°C.—J. Clay & F. van der Maesen. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 467–480. In English.) Experimental values for air, He, Ar and CO_2 are tabulated and discussed. A heterodyne beat method is used for determining the absolute values of small differences of capacitance by means of Clay's cylindrical measuring capacitor (141 below).

621.318.2 : 538.652 129

The Magnetostriction of Anisotropic Permanent Magnet Alloys.—M. McCaig. (*Proc. phys. Soc.*, 1st Oct. 1949, Vol. 62, No. 358B, pp. 652–656.) Blocks of permanent magnet alloys of the system FeNiAlCoCu were prepared with columnar crystals. The magnetostriction was measured in various directions before and after heat treatment in a magnetic field. The results agree well with those predicted theoretically, and so confirm the postulated crystal structure.

621.775.7 130

Sintering of Iron Powders.—H. Bernstorff. (*Metal Treatm.*, Summer 1948, Vol. 15, No. 54, pp. 85–89.) Tabulated results of experimental studies in the Degussa laboratories on the sintering of DPG Schleuder iron powder are used to derive curves showing tensile strength and elongation as a function of sintering time. For the higher sintering temperatures the curves approach constant values. At lower temperatures distinct minima are found for both strength and elongation.

666.1.037.5 131

Glass-to-Metal Seals.—(*Metal Ind., Lond.*, 30th Sept. & 7th Oct. 1949, Vol. 75, Nos. 14 & 15, pp. 263–266 & 292–293.) The relative merits of various metals and glasses used for sealing and various types of seal are discussed; characteristics of certain glasses are tabulated. For many purposes, B.T.H. C40 borosilicate glass sealed to Nilo-K, an alloy consisting of 54% Fe, 31% Ni and 15% Co, is the best combination now available. No great skill is required in preparing or manufacturing the seal; pre-beading is unnecessary. Seals with diameters up to at least 5 in. have been successfully made with these materials. The nature of stresses in seals is briefly discussed.

MATHEMATICS

512.974 : 621.3 132

Vectorial Space. Region of Representation for Electrotechnics.—G. Nasse. (*Bull. Soc. franç. Élect.*, Sept. 1949, Vol. 9, No. 96, pp. 459–474.) The concept of the region of representation, limited in classical teaching to the Fresnel plane, is shown to be capable of a natural extension leading to the vectorial space of geometers, a metric space capable of representing under one and the same structure the properties of monophase and polyphase systems, and even those of their transitory states.

681.142 133

The University of Manchester Universal High-Speed Digital Computing Machine.—T. Kilburn. (*Nature, Lond.*, 22nd Oct. 1949, Vol. 164, No. 4173, pp. 684–687.) The binary system is used, and numbers up to 2^{40} can be accommodated. Operations which the machine can perform and "addresses" where numbers can be stored

also have numbers. 'Instructions' take the form of numbers which specify that a certain operation is to be performed on the contents of a certain address. An electronic storage unit with a capacity of 5 120 digits is included; this type of storage was discussed in 2258 of 1949 (Williams & Kilburn). A magnetic storage unit is also available in which the numbers stored are less accessible but which has a capacity of 40 960 digits; a similar storage system was described by Booth (3179 of 1949). Special circuits are provided to perform addition, subtraction, multiplication and certain logical operations, but a 'sub-programme' is used for other operations, including division, which occur less frequently. The component elements of the computing circuits used are similar to those used in video pulse circuits in radar (see 1874 of 1948). One type of adding circuit is considered in detail. The 'control' is an electronic storage tube, holding two numbers; its operation is described. Arrangements are made to enable the machine to perform or omit an instruction according to the nature of the result obtained at a particular stage of the calculation.

681.142 : 621.385.032.212 **134**

Polycathode Glow Tube for Counters and Calculators.—Lamb & Brustman. (*See* 266.)

MEASUREMENTS AND TEST GEAR

531.764.5 **135**

The Development of the Quartz Clock.—W. A. Marrison. (*Elektrotechnik, Berlin*, Oct. 1949, Vol. 3, No. 10, pp. 311–316.) Condensed version of the paper noted in 762 of 1949.

621.317.2 **136**

New High-Voltage Engineering Laboratory.—J. H. Hagenguth. (*Gen. elect. Rev.*, Sept. 1949, Vol. 52, No. 9, pp. 9–14.) The laboratory was built for development and research work on apparatus used for power transmission and distribution. Test power for a.c. h.v. tests is supplied by one 1 000-kVA generator and two 500-kVA generators. Two 500-kW 500-V d.c. generators are also available. For studying transient voltages, such as those due to lightning, under controlled conditions, two 5 100-kV impulse generators are provided. These can produce a peak voltage of 7 500 kV to ground at a current of 33 kA when operated individually, or 66 kA when connected in parallel. Five 350-kV 1 000-kVA test transformers can be connected in cascade to deliver 1 750 kV (r.m.s.) to ground. A surge current generator is also available, capable of producing 260 kA at 50, 100 or 150 kV.

621.317.3.011.5 : 621.365.55† **137**

Dielectric Loss with Changing Temperature.—J. B. Whitehead & W. Rueggeberg. (*Elect. Engng, N.Y.*, Oct. 1949, Vol. 68, No. 10, p. 874.) Summary only. A method of measuring the properties of a dielectric while its temperature is changing has been developed and applied to certain thermosetting materials subjected to h.f. heating. The presence of absorbed moisture in small amounts has a pronounced effect on the dielectric properties of such materials. The two important phenomena are dielectric absorption and some form of molecular polarization; their relative importance depends on the type of heating cycle.

621.317.3.029.3 **138**

Audio-Frequency Measurements.—W. L. Black & H. H. Scott. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1108–1115.) Discussion of the theory involved in making measurements of gain, frequency response, distortion and noise, with particular reference to high-gain systems. Techniques of measurement and factors affecting accuracy are also discussed. See also 147 and 148 below.

621.317.32 **139**

Problems Related to Measuring the Field Strength of High-Frequency Electromagnetic Fields.—R. Truell. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1144–1147.) Continuation of 429 of 1949, with more detailed treatment of the parallel-plate field and extension to the field in a rectangular cavity. In both cases the field amplitude can be expressed quite simply in terms of easily measurable parameters, if certain velocity conditions are satisfied by the electron beam, and the cyclotron frequency $|e|H/mc$ is equal to the frequency of the e.m. field.

621.317.32 : 551.594 **140**

A New Method of Measuring the Vertical Electric Field in the Atmosphere.—J. A. Chalmers. (*J. sci. Instrum.*, Sept. 1949, Vol. 26, No. 9, pp. 300–301.) A metal ball rolls down an inclined earthed gutter and drops into an insulated collector, which thus acquires a potential proportional to the atmospheric electric field. This potential is amplified and measured by a galvanometer.

621.317.335.2† : 621.319.4 **141**

Accurate Determination of the Absolute Capacity of Condensers : Part 2.—J. Clay. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 484–488. In English.) Results are given of new precision measurements of the dimensions of three variable steel capacitors, from which their capacitances are deduced. Values obtained agree to within 1 part in 3000 with those found by e.s. and oscillation methods, and the absolute value is believed to be accurate within 0.01%. For part 1 see 4175 of 1936. These capacitors were used for the measurements of dielectric constant discussed in 125 and 128 above.

621.317.37.029.6 : 621.315.612 **142**

The Measurement of the Dielectric Properties of High-Permittivity Materials at Centimetre Wavelengths.—Powles & Jackson. (*See* 121.)

621.317.372 + 621.317.334 **143**

***Q* and L Measurements.**—W. T. Cocking. (*Wireless World*, Nov. 1949, Vol. 55, No. 11, pp. 449–453.) When a *Q* meter is not available, the frequency or capacitance detuning methods here described form convenient ways of determining *Q* and also the inductance and self-capacitance of coils; *Q* is obtained by calculations involving measured values of frequency or capacitance. Factors limiting accuracy are discussed.

621.317.372 : 621.315.614 **144**

Electrical Testing of Capacitor Paper.—H. S. Endicott. (*Gen. elect. Rev.*, Sept. 1949, Vol. 52, No. 9, pp. 28–35.) Equipment is described for measuring the dielectric constant and power factor of dry, unimpregnated capacitor paper. The sample is not removed from the drying chamber. The relation between the power-factor/temperature curves of dry and impregnated paper is discussed. The time necessary for a test has been reduced from 3 weeks to 2 days.

621.317.7 + 621.38 + 621.396.69] : 061.4 **145**

Radiolympia Review.—(*Wireless World*, Nov. 1949, Vol. 55, No. 11, pp. 428–444.) An illustrated review of various exhibits, namely (*a*) television equipment, (*b*) broadcast receivers, (*c*) sound-reproducing equipment, (*d*) components and accessories, (*e*) communication equipment, (*f*) testing and measuring gear, (*g*) scientific, industrial and medical apparatus. See also 3472 of 1949.

621.317.7.001.4 **146**

Operation and Care of Circular-Scale Instruments: Parts 1 & 2.—J. Spencer. (*Instruments*, July & Aug. 1948, Vol. 21, Nos. 7 & 8, pp. 620–632 & 715–724.) A fully-illustrated discussion of the mechanism of such instruments, and dismantling and reassembling

procedure for repair, showing the principles of operation of Westinghouse and General Electric moving-coil d.c. meters Types KX24 and DB-12, and moving-iron a.c. meters Types KA24 and AB-12. Calibration adjustments, maintenance, and repair of various defects, from simple frictional defect to component failure, are also considered.

621.317.7.029.3 147

Audio Frequency Measurements: Part 1.—W. L. Black & H. H. Scott. (*Audio Engng, N.Y.*, Oct. 1949, Vol. 33, No. 10, pp. 13–19 ; 43.) Paper presented at a joint I.R.E./R.M.A. meeting at Syracuse, N.Y. Definitions and minimum standards for the audio facilities of a broadcasting system are specified in R.M.A. Standard TR-105A, published in May 1948. The technical background which is the basis for this specification is here surveyed and measurement methods are discussed thoroughly. A test circuit for gain measurements is described. The use of calibrated adjustable attenuators is recommended. Sources of error include mismatch and variation in the degree of earthing. The measurement of bridging gain is considered. Part 2: 148 below.

621.317.7.029.3 148

Audio Frequency Measurements: Part 2.—W. L. Black & H. H. Scott. (*Audio Engng, N.Y.*, Nov. 1949, Vol. 33, No. 11, pp. 18–19 ; 50.) Continuation of 147 above. Determination of harmonic distortion at the output, and intermodulation and noise measurements are discussed. See also 138 above.

621.317.7.088.22 149

The Ratio of Electrical to Friction Torques in Indicating Instruments.—G. Szabó. (*J. sci. Instrum.*, Sept. 1949, Vol. 26, No. 9, pp. 301–304.) Factors affecting this ratio at its most favourable value are discussed. The shock-acceleration and shock-velocity are computed for the range of deformations where pivots and jewels are not damaged.

621.317.71 : 621.385.5 150

The Use of Multigrid Tubes as Electrometers.—J. R. Prescott. (*Rev. sci. Instrum.*, Aug. 1949, Vol. 20, No. 8, pp. 553–557.)

621.317.715 151

The Galvanometer as a Built-In Component.—G. Spiegel. (*Elect. Mfg, N.Y.*, July 1948, Vol. 42, No. 1, pp. 80–83.) Both the jewelled and the taut-suspension types of galvanometer are considered. The latter have much greater sensitivities and longer periods.

621.317.72 152

Compensator — A Device for Measurement of Alternating Currents within a Broad Frequency Range.—V. Hlavsa. (*Tesla tech. Rep., Prague*, March 1949, pp. 4–10.) For measurements at frequencies between 20 c/s and 20 kc/s. One of two synchronous generators supplies current to the object under test and the voltage developed across it is balanced by the output of the other generator. The magnitude and phase of this output voltage are controlled by means of a potentiometer and a phase shifter, both of which are calibrated. Design details and applications of the equipment are discussed.

621.317.726 : 621.396.615.17 153

Measurement of the Time Constants of Peak Voltmeters intended for the Study of Transient Phenomena.—F. de Clerck. (*HF, Brussels*, 1949, No. 3, pp. 81–87.) For measuring the time constant of the indicating instrument, a rectangular pulse is used of duration variable and of the order of the time to be measured. For the time of charge of the electrical circuit a sine-wave pulse, symmetrical with respect to zero and with a rectangular envelope, is used; the pulse duration is again variable. The discharge time requires the sudden interruption, for a controlled time, of a sine wave. A generator for producing these three types of pulse is described and also a method of using it to enable the times to be measured on the screen of a c.r.o. Other applications of the equipment are mentioned.

621.317.729 : 532.517.2 154

Fields from Fluid Flow Mappers.—A. D. Moore. (*J. appl. Phys.*, Aug. 1949, Vol. 20, No. 8, pp. 790–804.) Fluid flow within the streamline range, made visible by means of potassium permanganate crystals, can be used to simulate e.s., e.m. and other fields. Flow takes place between a flat slab of plaster-of-Paris, provided with suitable sources and sinks, and a parallel sheet of plate glass. Setting-up techniques for particular problems are described. Two-dimensional fields can in general be set up easily; some symmetrical three-dimensional fields can also be simulated. Photographs of results obtained are included and discussed.

621.317.73 155

Impedance Measurements with Directional Couplers and Supplementary Voltage Probe.—B. Parzen. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1208–1211.) The impedometer consists of two oppositely connected directional couplers and a voltage probe in a short transmission line. An experimental model for frequencies between 50 and 500 Mc/s is described.

621.317.733 156

The Design and Construction of a Comparison Impedance Bridge for Frequencies of 40-270 Mc/s.—W. C. Weatherley. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 429–432.) Two types of bridge are described, one for measuring 'balanced' and the other for 'unbalanced' impedances to earth. The frequency range is about ± 20% of the working mid-frequency. Measurements can be obtained without placing bulky apparatus near the impedance being measured. Changes in impedance over a definite frequency band are measured, rather than absolute impedance. Circuit details, suitable components and calibration procedure are described. Accuracy is within ±2% for resistance comparison and within ±5% for capacitance comparison.

621.317.74 : 621.395.44 157

Transmission-Measuring Set for Low-Frequency Carrier Systems.—J. Brundage & J. Zyda. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, pp. 204–208.) Description of the Type 902-A set, which includes a stabilized RC oscillator covering the frequency range 300–40 000 c/s. A 4-position rotary switch provides for sending, receiving, measuring and calibration. Transmission power may be varied from −40 to +20 db relative to 1 mW, and receiver sensitivity is adequate. Self-calibration is arranged. The unit may be either of portable form or rack-mounted.

621.317.75 158

Harmonic Analyzer and Synthesizer.—J. Lehmann. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 106–110.) Any waveform which can be represented by a Fourier sine or cosine series having up to 20 terms can be analysed by this electromechanical instrument. Two dials are provided on which the amplitude and phase of each term can be set. For synthesis, the amplitudes and phases of the first 20 odd harmonics are set on the dials, and the response is plotted on a dynamometer-type recorder. For analysis, the amplitudes of the steps of a stepped-wave approximation to the waveform under investigation are set on the amplitude dials, and the frequency response is obtained as two plotted curves from twin recorders, in rectangular or polar coordinates. Circuit arrangements are discussed, and examples illustrating the operation of the instrument are included.

621.317.755 : 621.385.029.63/.64 159

Traveling-Wave Oscilloscope.—Pierce. (*See* 264.)

621.317.761 160

Electronic Frequency Meter.—(*J. sci. Instrum.*, Sept. 1949, Vol. 26, No. 9, p. 310.) Made by Airmec Laboratories Ltd, High Wycombe. Designed to measure frequencies 0–200 c/s, 0–2 kc/s and 0–20 kc/s. Input signal can be between 0.1 and 20 V; the wave may be of any shape having not more than 2 zeros per cycle. A circuit diagram is given showing the waveform at various stages. Pulses of recurrence frequency corresponding to the signal frequency are derived. The meter indicates the number of pulses received per second. Accuracy within 1% is unaffected by mains voltage fluctuations not exceeding 10%.

621.317.761 161

A New Frequency-Meter and its Application.—H. Hochrainer. (*Elektrotech. u. Maschinenb.*, Oct. 1949, Vol. 66, No. 10, pp. 288–291.) The current whose frequency is to be determined passes through a barretter and thence through a capacitor and a resistor respectively to *opposite corners* of a rectifier bridge. If the currents through the capacitor and resistor are equal, a moving-coil ammeter connected across the other diagonal will show no deflection. If, however, the frequency alters, the capacitive current will change and a deflection will be obtained on one side or the other of the instrument's centre zero. Applications to frequency recording for power systems and to frequency control are considered.

621.317.763 : 621.392.26† 162

Wave-Guide Interferometers as Differential Wave-Meters.—A. B. Pippard. (*J. sci. Instrum.*, Sept. 1949, Vol. 26, No. 9, pp. 296–298.) The advantages of non-resonant waveguide interferometers over resonant cavities as absolute differential wavemeters are discussed, and two simple interferometers are considered in detail. An interferometer using a hybrid junction, and operating similarly to Michelson's interferometer, can have extremely high sensitivity in detecting small changes of frequency; experimental results are discussed.

621.317.772 163

On a Valve-Type Phase Meter.—J. Gilbert. (*Helv. phys. Acta*, 15th Aug. 1949, Vol. 22, No. 4, pp. 409–411. In French.) Description, with circuit diagram giving no component details, of a meter for reading directly the phase difference between two l.f. voltages of any waveform, and any amplitude between 20 V and 250 V. For frequencies in the range 20–500 c/s the accuracy is of the order of $\pm 2°$. A phase meter based on similar principles has been described by Florman & Tait (1726 of 1949). See also 3483 of 1949 (Kretzmer).

621.317.79 : 621.396.615 164

Citizens' Band Signal Generator.—W. C. Hollis. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 77–79.) Construction details of a unit which contains a tunable concentric-line resonator in a Colpitts oscillator circuit using a subminiature valve. For earlier articles on Citizens' Radio see 3482 of 1949 (Lurie) and back references.

621.317.79 : 621.396.822 165

A Video-Frequency Noise-Spectrum Analyzer.—P. S. Jastram & G. P. McCouch. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1127–1133.) Requirements for noise-analyser design are discussed. The heterodyne method of analysis is preferred to the tuned-circuit method because wide frequency coverage can be obtained with constant sensitivity and simple tuning arrangements. Design procedures can be worked out for obtaining adequate dynamic operating range and for suppression of errors. All circuits after the first modulator operate over fixed narrow frequency bands. A practical instrument for a frequency range of 50 kc/s to 10 Mc/s is described. The width of the analysing pass band is 33 kc/s.

621.317.79 : 621.396.933 166

Distant Monitoring of Radio Transmissions for Naval and Aviation Services.—J. Marique. (*HF, Brussels*, 1949, No. 3, pp. 71–80.) The subject is considered as affecting ship or aircraft security, navigation and communications. Examples illustrate the results of measurements carried out at the special monitoring centre established in Brussels since 1946. An outline is given of the methods adopted there for the measurement of frequency, the recording of signals in a given frequency band and analysis of the records.

621.317.794 167

A Ten Centimeter Broadband Bolometer Cavity.—T. Miller. (*Tele-Tech*, Sept. 1949, Vol. 8, No. 9, pp. 28–31, 74.) A 10-mA fuse is used as the bolometer element of a cavity resonator for measuring powers from 100 μW up to several milliwatts. Power reflection is $<1\%$ for wavelengths between 9 and 10.5 cm. The cavity bridge circuit and methods of increasing bandwidth by means of a resonant diaphragm soldered into the waveguide are described.

621.319.4.089.6 168

The Calibration Curves for Kohlrausch Capacitors.—G. Zickner. (*Elektrotechnik, Berlin*, Nov. 1948, Vol. 2, No. 11, pp. 317–320.) Description of a parallel-plate capacitor with variable plate distance, and discussion of the representation of its calibration curve in cartesian, double-logarithmic or hyperbolic coordinates, and also of the best means for determining the error curve showing the departure from the exact hyperbolic law.

621.385.001.4 : 621.3.015.3 169

Surge Testing of High Vacuum Tubes.—H. J. Dailey. (*Tele-Tech*, Oct. 1949, Vol. 8, No. 10, pp. 26–29, 60.) An experimental investigation with artificially generated flash-arcs. Results indicate that (*a*) adding series resistors in the anode circuit does not materially affect valve damage per flash-arc for low gas pressures, but does reduce the tendency towards arcing, (*b*) shielding the filament supports from the anode causes prompt failure, (*c*) internal sources of gas must be minimized, (*d*) initial gas pressure should be of the order 10^{-6} mm Hg to minimize flash-arc damage, (*e*) the filament exposed to flash-arc should be at maximum operating temperature, (*f*) the addition of a series resistor may increase flash-arc damage if the valves are soft and remain soft.

621.396.621.001.4 : 621.3.015.3 170

Transient Phenomena in Radio Receivers.—Carniol. (*See* 218.)

621.396.645.37.001.4 : 621.3.016.352 : 621.3.015.3 171

Examination of Amplifier Stability by applying a Sudden D.C. Test Voltage.—Carniol. (*See* 77.)

621.397.62.001.4 : 621.396.619 172

Measuring Modulation Depths of TV Signals.—R. P. Burr. (*Tele-Tech*, Sept. 1949, Vol. 8, No. 9, pp. 32–35, 77.) Description of methods used and equipment required, especially for cases where only a small test signal is available. See also 3593 of 1947 (Buzalski).

621.317 173

Hochfrequenzmesstechnik. [Book Review]—O. Zinke. Publishers: S. Hirzel, Leipzig, 2nd edn 1947, 253 pp. (*Elektrotechnik, Berlin*, Nov. 1948, Vol. 2, No. 11, p. x.) There are twelve main sections dealing with current, voltage, field-strength and power measurements and with practically every type of measurement required in radio technique. "... an extraordinarily valuable help for every h.f. technician."

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.715.4 : 621.396.619.13 174

Measurement and Amplification of Small Displacements by means of F.M.—P. Bricout & M. Boissvert. (*Rev. gén. Elect.*, Oct. 1949, Vol. 58, No. 10, pp. 402–404.) An instrument is described in which small displacements of the plate of a specially-contructed capacitor in an oscillatory circuit produce frequency variations at the mixer grid of a heptode in a Hartley circuit. This discriminator is due to Bradley (227 of 1947) and is relatively insensitive to amplitude variations; it operates linearly over a large part of the scale. Sensitivity is adjustable within wide limits; amplification is above 1 000 and relative error below 1%. Applications of the instrument in various pressure measurements are noted.

534.321.9.001.8 : 539.61.08 175

Evaluation of Adhesion by Ultrasonic Vibrations.—S. Moses & R. K. Witt. (*Industr. Engng Chem.*, Oct. 1949, Vol. 41, No. 10, pp. 2334–2338.) A direct, quantitative method for measuring the adhesion of organic coatings to both metallic and non-metallic surfaces is described. Longitudinal ultrasonic vibrations are induced by an electrodynamic system in a metal cylinder, one end of which is threaded for attachment of the sample to be coated. The coating separates from the face of the sample when the force required to accelerate it exceeds the adhesion force at the interface. The accelerating force is determined by the frequency and amplitude of the vibration, and the mass and dimensions of the film.

535.247.4 176

Darkroom Light Meter.—S. Becker. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 90–91.) A sensitive photocell unit samples the light falling upon an enlarging easel and gives the optimum exposure time directly in seconds. Complete construction details are included.

535.61-15 : 621.385.832 177

Cathode-Ray Presentation for Infrared Spectrometer.—J. H. Jupe. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 189–193.) Description of an instrument capable of scanning a band of width 3 μ anywhere in the range 2–16 μ, in a total time of about 15 sec.

539.16.08 178

Electron Mobilities in Geiger-Müller Counters.—H. den Hartog, F. A. Muller & C. S. W. van Rooden. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 581–587. In English.)

539.16.08 179

Geiger Tube Quenching Circuit for a Negative High Voltage Supply.—R. J. Watts. (*Rev. sci. Instrum.*, Sept. 1949, Vol. 20, No. 9, p. 689.) A modified Neher-Harper circuit. The Type CK5694AX valve was selected for a portable design; other heater types of valve can be used.

539.16.08 180

Speed of Operation of Geiger-Müller Counters.—H. den Hartog. (*Nucleonics*, Sept. 1949, Vol. 5, No. 3, pp. 33–47. Bibliography, p. 47.) Discussion of the delays caused by electron transit time, discharge development, discharge quenching and dead-time. The work of various authors who have attempted to explain or measure some of the complex phenomena involved is reviewed. The resolution of G-M counters is compared with that of other counters.

539.16.08 : 621.386 181

Dead-Time and Non-Linearity Characteristics of the Geiger-Counter X-Ray Spectrometer.—L. Alexander, E. Kummer & H. P. Klug. (*J. appl. Phys.*, Aug. 1949, Vol. 20, No. 8, pp. 735–740.)

621.316.7.076.7 : 543.712 : 676/677 182

Automatic Control of Moisture.—R. V. Coles. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 82–86.) Electronic equipment makes corrections proportional to the difference from the desired moisture content as material passes through a capacitive measuring element. This type of control is specially useful in the textile and paper industries.

621.317.083.7 183

Electrical Telemetering.—N. A. Abbott. (*Radio & Televis. News, Radio-Electronic Engng Supplement*, Oct. 1949, Vol. 13, No. 4, pp. 3–6, 29.) Discussion of primary detectors for the actual measurements, intermediate links for transmitting the information, and indicating or recording units. A block diagram of a typical complete telemetering system is included.

621.317.39 : 531.775 184

A Tachometer.—A. E. Benfield. (*Rev. sci. Instrum.*, Sept. 1949, Vol. 20, No. 9, pp. 663–667.) Theory and operation of a simple instrument in which a d.c. e.m.f. is generated by means of a magnet attached to the rotating member. A formula for the e.m.f. is given, and the effect of varying the position and orientation of the magnet is considered. The e.m.f. is directly proportional to the speed of rotation, but is rather small; possible ways of increasing it are briefly discussed.

621.317.39 : 620.178.3 + 531.768 185

Modern Vibration Meters with Electrical Indication.—H. Köhler. (*Elektrotechnik, Berlin*, Oct. 1949, Vol. 3, No. 10, pp. 301–310.) The physical principles of vibration and acceleration meters are considered and short descriptions are given of a wide variety of instruments, including capacitive, electrodynamic and piezoelectric types. These fall into two groups: (*a*) with relatively low sensitivity for investigations dealing with machinery; (*b*) with high sensitivity for use in seismology and in building and mining research. A fuller account is given of a piezoelectric acceleration meter with very high sensitivity; this has a longitudinally stressed Rochelle-salt crystal as indicating element. Results obtained with this meter are described.

621.365.5 186

Coupling Circuits for H.F. Heating.—R. A. Whiteman. (*Radio & Televis. News, Radio-Electronic Engng Supplement*, Oct. 1949, Vol. 13, No. 4, pp. 8–11, 27.) Design of such circuits for optimum power transfer and optimum loading of the power stage.

621.365.54† 187

(Vacuum) **Induction Furnace for High-Temperature Ceramic Research.**—P. D. Johnson. (*J. Amer. ceram. Soc.*, Oct. 1949, Vol. 32, No. 10, pp. 316–319.) An illustrated description. Power is supplied by a 9 600-c/s generator driven by a 90-h.p. 3-phase motor. Temperatures up to 2 300°C can be obtained by using Mo for the susceptor and the radiation shields.

621.365.54† : 621.785.6 188

Multi-Purpose Inductance Hardening Units.—(*Machinery, Lond.*, 22nd Sept. 1949, Vol. 75, No. 1926, pp. 413–415.)

621.365.55† 189

Temperature Behaviour in Electrically Heated Non-homogeneous Bodies.—W. F. Kussy. (*Elektrotechnik, Berlin*, Oct. 1949, Vol. 3, No. 10, pp. 323–326.) Calculations with particular application to porcelain coil-formers, for both short-period and continuous loading.

621.365.55† : 621.317.3.011.5 190

Dielectric Loss with Changing Temperature.—Whitehead & Rueggeberg. (*See* 137.)

621.38.001.8 191

Electronics in War and Peace.—A. L. Whiteley. (*Engineering, Lond.*, 30th Sept. 1949, Vol. 168, No. 4366, pp. 350–352.) Long summary of paper read before the British Association. The advantages of electronics for war purposes are high sensitivity and high speed of response. Industrial applications to welding and printing are discussed. The development of electronics in industry has been slower than expected because the installation of electronic devices often requires the re-design of machinery and the training of operating and maintenance staffs. The accuracy of electronic devices is often far greater than that required. Valve life must be lengthened. Electronic engineers must have a closer understanding of the processes they hope to improve.

621.38.001.8 : 629.13.052 192

An Electronic Pressure-Sensitive Transducer.—G. Day. (*J. sci. Instrum.*, Oct. 1949, Vol. 26, No. 10, pp. 327–329.) The applied pressure varies the interelectrode spacing of a double-diode valve working under space-charge-limited conditions. The resultant changes in the two anode currents produce an out-of-balance voltage in a bridge circuit of which the diodes form two of the arms. The device is primarily designed for use as an aircraft altimeter.

621.383.001.8 : 535.61-15 193

Applying the Infrared Image Converter Tube.—R. D. Washburne. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 150–164.) Description of a British tube, Type CRI-143, in which conversion is achieved by means of a uniform field between the anode and the cathode. The image is not inverted, and the tube is much simpler than its American counterpart. For another account see 1710 of 1948 (Pratt).

621.383.001.8 : 536.587 194

Light-Sensitive Cells.—(*Overseas Engr*, Nov. 1949, Vol. 23, No. 265, pp. 128–129.) Applications in industry are discussed, with particular reference to automatic temperature control during welding and pre-heating operations.

621.384.611† 195

Betatron Injection into Synchrotrons.—F. K. Goward. (*Proc. phys. Soc.*, 1st Oct. 1949, Vol. 62, No. 358A, pp. 617–631.) Factors influencing the proportion of electrons which may be trapped and accelerated in a transition from betatron to synchrotron operation are discussed; this transition is important in synchrotron design. Machines of various energies are studied; satisfactory agreement between theory and experimental results is obtained.

621.384.611.2† 196

A 30-Million Volt Synchrotron for Medical Use.—(*Nature, Lond.*, 29th Oct. 1949, Vol. 164, No. 4174, pp. 726–728.) A fuller description of the equipment noted in 3223 of 1949 (Martin).

621.384.611.2† 197

The Dynamics of a Synchrotron with Straight Sections.—N. M. Blachman & E. D. Courant. (*Rev. sci. Instrum.*, Aug. 1949, Vol. 20, No. 8, pp. 596–601.)

621.384.612.1† 198

The Electron Cyclotron.—W. J. Henderson & P. A. Redhead. (*Nucleonics*, Oct. 1949, Vol. 5, No. 4, pp. 60–67.) An experimental model, built by the National Research Council of Canada, is described. Magnetic field and applied r.f. are constant; the electron time lag per revolution due to the relativistic increase in mass is made equal to an integral number of periods of the r.f. field. This type of accelerator may have distinct advantages over other types in the energy region of 10^7–10^8 eV.

621.385.833 199

Electron-Optical Shadow Method.—(*Tech. Bull. nat. Bur. Stand.*, Sept. 1949, Vol. 33, No. 9, pp. 106–108.) A technique based on extensive theoretical analysis and developed at the National Bureau of Standards by L. L. Marton; theoretical formulae have been derived by S. H. Lachenbruch. An electron lens system is used to produce a shadow image of a fine wire mesh placed in the path of an electron beam. From the distortion in the shadow pattern caused by deflection of the electrons as they pass through the field under investigation, accurate values of field strength can be computed. The method can be applied to fields of very small dimensions, such as the fringe fields from the small domains of spontaneous magnetization in ferromagnetic materials. The method is somewhat similar to the electron-optical schlieren method, but is much better adapted to precise determination of field intensity.

621.385.833 200

A New Type of Focusing in a Magnetic Lens Field.—H. Slätis & K. Siegbahn. (*Phys. Rev.*, 15th June 1949, Vol. 75, No. 12, p. 1955.)

621.396.615 : 621-12 201

The Reciprocator.—White & Lord. (*See* 66.)

PROPAGATION OF WAVES

535.42 : 538.56 202

On the Diffraction of an Electromagnetic Wave through a Plane Screen.—J. W. Miles. (*J. appl. Phys.*, Aug. 1949, Vol. 20, No. 8, pp. 760–771.) The screen is assumed infinitely thin and perfectly conducting. The diffraction problem is formulated, using generalized cylindrical coordinates, in terms of the generalized Fourier transform of the tangential electric field in the aperture. An integral equation for this transform is obtained. The power transferred through the aperture is calculated and expressed in a variational form of the Schwinger type. The significance and behaviour of the aperture impedance is considered. Other formulations of the problem, including one involving Babinet's principle (1335 of 1947), are discussed. The variational formulation appears to provide a convenient link between the results for large wavelengths for which Rayleigh's static methods are valid, and those for small wavelengths where geometrical optics can be used. It appears to be superior to the Kirchhoff theory for any assumed aperture field. See also 1845 of 1948 (Levine & Schwinger).

535.42 : 621.396.81.029.6 203

Diffraction of High-Frequency Radio Waves around the Earth.—M. D. Rocco & J. B. Smyth. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1195–1203.) The results of height-gain measurements at 7 frequencies in the range 25–9 375 Mc/s for a nonoptical path in the Gila valley, Arizona, can be adequately explained by diffraction theory alone for the lower frequencies. Fields observed at the higher frequencies for low terminal heights were considerably stronger than those predicted by standard diffraction theory, even under meteorological conditions which were subnormal near the ground. These strong fields had rapid variations with time but no variation with height. The frequency distribution of the signal fluctuations agrees with a Rayleigh distribution. Interpretation of the data according to the waveguide theory of atmospheric propagation (507 and 2892 of 1947) is being investigated.

538.566 204

Transmission of Electric Waves through Wire Grids.—W. Franz. (*Z. angew. Phys.*, June 1949, Vol. 1, No. 9, pp. 416–423.) The transmission of plane e.m. waves through a system of parallel grids of identical grid-constant is calculated, assuming that the radius of the wires is small with respect to the wavelength and to the wire spacing. The radiation resistances occur as coefficients in a system of linear equations from which the excitation of the individual grids and also the transmission properties of the whole system can be determined. Comparison with the measurements of Esau, Ahrens & Kebbel (2631 of 1939) shows satisfactory agreement.

538.566.2 : 535.13 205

The Propagation of Electromagnetic Waves through a Stratified Medium and its WKB Approximation for Oblique Incidence.—H. Bremmer. (*Physica, 's Grav.*, Aug. 1949, Vol. 15, No. 7, pp. 593–608.) The plane-wave solution of Maxwell's equations for a stratified medium is split into a series of terms, which have a simple physical meaning; the first of these terms constitutes the WKB approximation. By the introduction of a convenient Hertzian vector, the original vector problem is made scalar. The application of the saddle-point method to the individual terms of the series leads to simple geometric-optical approximations.

621.396.11 206

Velocity of Electromagnetic Waves.—C. I. Aslakson. (*Nature, Lond.*, 22nd Oct. 1949, Vol. 164, No. 4173, pp. 711–712.) The U.S. Air Forces used shoran extensively for measuring geodetic distances of from 67 to 367 miles. The results appeared to indicate that the velocity of e.m. waves was 299 792 km/sec. The method of observation was to determine the minimum of the sum of the distances between an aircraft and two ground stations as the aircraft flew across the line between the stations. The distances were corrected for the velocity of radio waves by a numerical integration along the ray path based on nearly simultaneous psychometric observations. See also 3488 of 1948 (Essen & Gordon-Smith) and 208 below.

621.396.11 : 551.510.535 207

Some Important Results of the Geometrical-Optical Properties of the Ionosphere.—K. Rawer. (*Rev. sci., Paris*, May/June 1948, Vol. 86, No. 3296, pp. 481–485.) The results are presented of calculations of the paths of rays reflected from an ionosphere layer with a parabolic ionization distribution, the angle of projection of the rays having any value up to 90°. Curves are also given for the field strength and the m.u.f. factor as a function of distance from the transmitter. Values calculated for the radius of the zone of silence for different angles of projection are in good agreement with the values found from formula (12) of Appleton & Beynon's paper noted in 3290 of 1940.

621.396.11.029.64 208

The Measurement of the Velocity of Propagation of Centimetre Radio Waves as a Function of Height above the Earth: Part 2 — The Measurement of the Velocity of Propagation over a Path between Ground and Aircraft at 10 000, 20 000 and 30 000 ft.—F. E. Jones & E. C. Cornford. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 447–452.) Continuation of 1441 of 1948 (Jones). Results of observations at two Oboe ground stations are given. The most probable values for the mean velocity of propagation between ground and aircraft are 299 713 km/s, 299 733 km/s and 299 750 km/s for aircraft heights of 10 000 ft, 20 000 ft and 30 000 ft respectively.

621.396.11.029.66 209

Propagation Characteristics of Tenth-Millimetre Waves.—R. Franz. (*Radio Tech., Vienna*, Aug. & Oct. 1949, Vol. 25, Nos. 8 & 10, pp. 461–464 & 581–583.) Absorption, transmission and selective-reflection properties of various solids for these waves are reviewed and the quartz-lens and residual-ray methods of isolating them are outlined. Absorption and reflection properties of gases and dipolar substances, and resonance phenomena in rock salt and other materials are considered.

621.396.81 : 621.397.5 210

WDTV Field-Strength Report.—Goldsmith. (*See* 250.)

621.396.812 + 538.566.3 211

A Survey of Ionospheric Cross-Modulation (Wave-Interaction or Luxembourg Effect).—L. G. H. Huxley & J. A. Ratcliffe. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 433–440.) Existing theoretical and experimental knowledge is surveyed. The pioneer theory of Bailey & Martyn (1934 Abstracts, p. 199 and p. 606) is restated in a form which relates it more closely to standard ionosphere theory. The experimental results are summarized in a form which enables the magnitude of the effect to be deduced approximately for any pair of stations. The way in which observations of cross-modulation can be used in ionosphere research is outlined. See also 1767 of 1949 (Cutolo & Ferrero).

621.396.812.029.62 212

Painless Prediction of Two-Meter Band Openings.—W. F. Hoisington. (*QST*, Oct. 1949, Vol. 33, No. 10, pp. 22–25.) Several examples of long ranges for amateur communication are correlated with the corresponding weather maps. Long ranges are associated with areas of high barometric pressure. The trailing edges of such high-pressure areas are particularly important.

621.396.812.3.029.56 213

Tropospheric Effects in Short and Medium Radio Wave Propagation.—W. J. G. Beynon. (*Nature, Lond.*, 22nd Oct. 1949, Vol. 164, No. 4173, p. 711.) The results obtained by Heightman (2308 of 1949) are confirmed by measurements made in July 1945 at Loth, Sutherland, on wavelengths between 60 and 300 m. Echo signals of appreciable amplitude appeared at ranges of 45 km and 65 km. They could be identified as reflections from mountains and were observed at all frequencies between 1 Mc/s and 5 Mc/s. Amplitude was often constant for several minutes or even hours, but sometimes varied by a factor of 4 or 5 to 1 in a few seconds. No well-defined variation in mean amplitude was found.

RECEPTION

621.396.621 + 621.397.62 214

Mass Production of Radio and Television Receivers.—M. Alixant. (*Radio tech. Dig., Édn franç.*, Aug. 1949, Vol. 3, No. 4, pp. 197–213.) Discussion of modern methods, from the prototype to the finished and tested article.

621.396.621 215

Features of French Broadcasting Receivers for the Season 1949-1950.—J. Rousseau. (*T.S.F. pour Tous*, Oct. 1949, Vol. 25, No. 252, pp. 325–328.) A list is given, with details of valves fitted in each receiver, type of circuit and number of stages. Only a few models have a h.f. stage preceding the frequency changer.

621.396.621 216

A 400-Mc/s Receiver Front End employing Subminiature Tubes and New Miniature Tuned Circuits.—V. H. Aske. (*Sylvania Technologist*, Oct. 1949, Vol. 2, No. 4, pp. 2–5.) A new type of matched single-tuned input circuit is used, with plunger-type tuning. This is

followed by a double-tuned r.f. stage and a pentode-type mixer working into a 30-Mc/s i.f. stage. The i.f. circuit is loaded with an impedance which simulates the input loading of the i.f. valve. Design considerations and circuit and performance details are discussed. Gain could be increased by reducing the bandwidth of the r.f. valves. Results show that valves are available which can be used satisfactorily for the awkward frequencies between 400 and 1 000 Mc/s.

621.396.621 : 621.317.35 217

On the Energy-Spectrum of an Almost Periodic Succession of Pulses.—G. G. Macfarlane. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1139–1143.) The energy/frequency spectrum is discussed for (*a*) regularly spaced pulses whose amplitudes have random oscillations about a mean value, (*b*) identical pulses whose recurrence rate varies in a random manner about a mean value. Both spectra have two components, a line spectrum and a continuous spectrum. In case (*a*), the envelope of each component is proportional to the envelope of the spectrum of a single pulse and the spacing of the lines in the line spectrum equals the repetition frequency. In case (*b*), the envelopes of the two spectra are not the same as that of a single pulse, and the spacing of the lines equals the mean repetition frequency. Extension of the method of calculation to other cases is briefly discussed.

621.396.621.001.4 : 621.3.015.3 218

Transient Phenomena in Radio Receivers.—B. Carniol. (*Tesla tech. Rep., Prague*, March 1949, pp. 21–34.) For testing the a.f. part of a receiver, square-wave oscillations with a relatively low fundamental frequency are used, so that any transients will terminate within less than half a period. Unstable frequencies well outside the a.f. range usefully transmitted may affect receiver performance considerably. For the h.f. stages, a rectangularly modulated h.f. signal is used. Relations between frequency characteristics, phase characteristics and transients are discussed. Transient phenomena occurring in tuned circuits and band-pass filters on the sudden application of a d.c. or h.f. voltage are considered in some detail and illustrated by oscillograms and phase characteristics. See also 77 above.

621.396.621.53 219

High Gain V.H.F. Converter.—H. O'Heffernan. (*Short Wave Mag.*, Feb. 1948, Vol. 5, No. 12, pp. 720–724.) Design, construction and component details for a 50–58-Mc/s unit of exceptional performance. The unit is arranged to have extremely short r.f. wiring. Each stage can be easily adjusted. Aerial coupling can be varied from the front panel. There is similar front-panel control of the first r.f. stage trimmer. Transformer or capacitive coupling is available between stages, and coils are easily accessible. The converter is designed to work into the common 1.6-Mc/s i.f. channel of the main receiver, and to be switched in when required.

621.396.621.54 220

New Reception Principle for Superheterodynes.—F. Tomek. (*Radio Tech., Vienna*, Oct. 1949, Vol. 25, No. 10, pp. 584–586.) Description of a patented circuit termed the "summadyne". Instead of selecting the difference between the signal frequency and that of the local oscillator as the receiver i.f., the sum of the two frequencies is chosen. In order, therefore, to keep the i.f. constant, the tuning of the local oscillator must be lowered as that of the input circuit is raised. This is effected by using identical components for the two circuits, with variable capacitors ganged in opposition. The advantages of such an arrangement are enumerated. A differential capacitor is used for tuning in two examples of receiver circuits, which both use double heterodyning and one of which has an amplifier stage for the first i.f. Intermediate frequencies of about 2 000 kc/s and 100 kc/s are used.

621.396.622 221

Signal-to-Noise Ratios of Linear Detectors.—R. H. DeLano. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1120–1126.) A practical approach to the problem of obtaining the signal and noise spectra of the output of a linear detector, given the input signal and the input noise power spectrum. Graphical Fourier analyses are performed since the corresponding analytical expressions are cumbersome. The ratio of the input bandwidth to the centre frequency is assumed small. Certain restrictions are assumed for the input signal: an a.m. signal, for example, is regarded as the product of a slowly varying envelope and a carrier-frequency sine wave. A comparison is made with square-law detection for a few useful cases. The linear detector gives a higher output signal/noise ratio than the square-law detector for some types of signal.

621.396.622 222

An Improved Synchronous Detector.—W. C. Michels & E. D. Redding. (*Rev. sci. Instrum.*, Aug. 1949, Vol. 20, No. 8, pp. 566–568.) The synchronous amplifier design of Michels & Curtis (43 of 1942) has been modified to minimize feedback and hence to allow pre-amplification and a sensitivity of 2×10^5 mm/V. The instrument has an inductive input impedance of 2 MΩ and a bandwidth of 0.25 c/s at 800 c/s. Its merits are discussed.

621.396.622 : 537.312.62 223

Detection at Radio Frequencies by Superconductivity.—J. V. Lebacqz, C. W. Clark, M. C. Williams & D. H. Andrews. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1147–1152.) Detection by superconducting CbN was studied as a function of r.f. current, bias current and temperature. A nonlinear resistance effect occurs in the transition region, with specially high values of dR/dI for currents < 1 mA. This is believed to be due to the effect of the magnetic field of the current superconductivity (Silsbee hypothesis). The observed values of dR/dI explain to a certain extent the observed rectified potentials at 1 Mc/s, but the increasing rectification observed at higher frequencies is not yet explained. See also 2531 of 1949 (Lebacqz & Andrews).

621.396.622 : 621.396.619.13 224

Experimental Tube for F.M. Detection.—L. J. Giacoletto. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 87–89.) A single-valve locked-in oscillator was used for f.m. detection by Bradley (227 of 1947) with a grid-controlled multigrid valve such as Type 6SA7. For optimum performance of this circuit, the current flowing beyond the second grid of the valve should not affect the oscillator circuit except through the feedback loop. In conventional multigrid valves, the oscillator grid and the input-signal grid are not sufficiently isolated to permit the use of a high-impedance input circuit. To overcome this difficulty a valve combining beam deflection and grid control was developed. Audio output on an early model was low, but proposed design changes may increase the average output current at least fivefold. Performance characteristics of various f.m. detectors are compared.

621.396.622.7 225

Limiter Discriminator versus Ratio Detector.—H. K. Milward & R. W. Hallows. (*Radio-Electronics*, Nov. 1949, Vol. 21, No. 2, pp. 20–22.) Results of tests carried out in England indicate that (*a*) although the difference in noise rejection is not great, the discriminator with a single limiter has slightly better performance than the ratio detector, (*b*) the discriminator with 2 limiters is

decidedly superior to the ratio detector, particularly when the noise level is high. Test apparatus and procedure are discussed. See also 2345 of 1948 (Maurice & Slaughter).

621.397.82 226

The Influence of U.H.F. Allocations on Receiver Design.—Reid. (*See* 258.)

STATIONS AND COMMUNICATION SYSTEMS

621.391.5 : 621.395.623.66 : 791.44 227

Inductive Prompting System.—B. H. Denney & R. J. Carr. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 66–69.) A system which allows the producer to communicate with actors in a film studio, and which does not interfere with the film sound system. A modulated magnetic field is detected and demodulated in receivers of the hearing-aid type. Receiver leads can be made photographically invisible or concealed under the actor's clothing. The receivers have neither valves nor batteries; crystal detectors are used. Each actor wears a coil of wire in which a secondary current flows; the 100-kc/s transmitter, of which circuit details are given, is connected to a single-turn loop which surrounds the set area and induces a strong r.f. field at all points within the area.

621.396.1 : 621.396.931 228

Radio Communications Services : Parts 2–4.—(*FM-TV*, July, Sept. & Oct. 1949, Vol. 9, Nos. 7, 9 & 10, pp. 18–21, 45, 21–24, 30, & 21–23.) Part 1: 2907 of 1949.

621.396.619 229

Neglected Outphasing System of Modulation.—W. H. Hartman. (*CQ*, Oct. 1949, Vol. 5, No. 10, pp. 18–20, 68.) Theory, design and constructional details of a practical transmitter using this form of modulation, which combines all the advantages of high-level modulation of a class-C amplifier with the economies of low-level systems. See also 85 of 1936 (Chireix).

621.396.619.16 230

A Method of Asymmetrical Pulse Duration Modulation.—R. J. Watts. (*Rev. sci. Instrum.*, Aug. 1949, Vol. 20, No. 8, pp. 622–623.)

621.396.65 : 621.396.931/.932 231

Review of the Applications of V.H.F. Radio Communications.—E. W. Northrop. (*G.E.C. J.*, Oct. 1949, Vol. 16, No. 4, pp. 184–196.) Discussion of typical equipment both for mobile and for point-to-point communication, choice of headquarters site, aerial height, optimum power and frequency, etc.

621.396.712 232

The "New Look" at KTBS.—(*Broadcast News*, Sept. 1949, No. 56, pp. 40–45.) General description of facilities available, with photographs of studios, control room etc.

621.396.931 233

The Planning of 'Business-Radio' Services at Very High Frequencies.—(*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 381–382.) Report on I.E.E. Radio Section discussion meeting. Two main types of service are considered: (*a*) an extension of the ordinary telephone service to mobile units, (*b*) an exclusive service for the various mobile units within one organization, such as a taxi company.

621.396.97 234

25 Years Broadcasting in Austria.—W. Füchsel. (*Radio Tech., Vienna*, Oct. 1949, Vol. 25, No. 10, pp. 576–580.) A review of developments at the R.A.V.A.G. station, Vienna. The 120-kW transmitter installed in 1932 on the Bisamberg, the 100-kW Graz-Dohl transmitter and low-power transmitters in Kärnten and Steiermark are mentioned. Subscribers now number over 1 300 000.

SUBSIDIARY APPARATUS

621-526 235

Servomechanisms and Modern Physics.—R. Moch. (*Radio tech. Dig., Édn franç.*, June & Aug. 1949, Vol. 3, Nos. 3 & 4, pp. 133–145 & 235–249...252.) The construction and operation of servomechanisms is studied and also the electronic or electromechanical calculating elements which can be incorporated in such devices, including differentiators, integrators and apparatus for other types of mathematical operations. Various examples of the application of servomechanisms are briefly described; these include process control, stabilization (as in the 'atomic clock' of the National Bureau of Standards) and calculating machines.

621-526 236

Design Equations for Servomechanisms.—B. Parzen. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, pp. 249–256.) Reprinted from the book noted in 268 below. Fundamental quantities, and relations between them, are derived for linear lumped-constant servomechanisms; positioning systems using electronic and electromechanical devices are considered in more detail. General theory and its application to typical positioning systems are discussed. Performance and stability criteria are briefly mentioned. See also 232 of 1949 (James, Nichols & Phillips).

621-526 237

Magnetic Fluid Clutch in Servo Applications.—G. R. Nelson. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 100–103.) Report on experience obtained with various iron-disk rotor designs running in a mixture of oil and powdered iron which solidifies when a magnetic field is applied. Such clutch units are useful in servomechanisms at natural frequencies below 30 c/s.

621.3.076.7 : 621.3.016.1 238

Torque and Speed Regulation with the Electronic Amplidyne.—J. L. Dutcher. (*Elect. Mfg, N.Y.*, July 1948, Vol. 42, No. 1, pp. 84–89...158.) Principles of the amplidyne, which differs from a conventional generator only in having an extra pair of short-circuited brushes, are discussed. Circuit diagrams showing its use for various applications are included.

621.3.077.2/.3 239

The Amplidyne Generator — Its Performance and Design.—M. S. Hoffenberg. (*Trans. S. Afr. Inst. elect. Engrs*, Aug. 1949, Vol. 40, Part 8, pp. 175–188. Discussion, pp. 188–191.) Discussion of: (*a*) the desirability of high amplification and low time constant in control apparatus, (*b*) the cross-field principle of operation of the amplidyne, its steady-state operation on d.c., and the effect of varying the compensation of the output stage, (*c*) the transient and steady-state response to a.c. signals of both perfectly and imperfectly compensated machines, (*d*) the effects of rocking the brushes and chording the armature windings, (*e*) two types of stator construction, (*f*) the use of an output coefficient as an aid to design, (*g*) the effect of field form factor on the average flux density, and of quadrature current on specific electric loading, (*h*) variation of power-amplification ratio with number of cycles, (*i*) the magnetic circuit and commutation, (*j*) the calculation of leakage fluxes and stage inductances.

621.314.6 : 621.315.59 240

Semiconductor Rectifiers.—(*Elect. Engng, N.Y.*, Oct. 1949, Vol. 68, No. 10, pp. 865–872.) Long summary, compiled by S. J. Angello, of 4 papers read at an A.I.E.E. symposium on 'Electrical Properties of Semiconductors and the Transistor'; namely:—Theory of Rectification, by F. Seitz. Boundary Layers in Rectifiers, by H. Y. Fan. Noise in Semiconducting Contacts, by P. H. Miller, Jr.

A Comparison between the Schottky Rectifier Theory and Measurements upon Cuprous Oxide Cells, by S. J. Angello.

241

621.314.63

High Inverse Voltage Germanium Rectifiers.—S. Benzer. (*J. appl. Phys.*, Aug. 1949, Vol. 20, No. 8, pp. 804–815.) Current/voltage characteristics were determined for various point-contact Ge rectifiers. Inverse voltages of several hundred volts were observed; a reproducible negative differential-resistance region occurs in the inverse characteristic. The metal used for the point contact has little effect; the effects of impurities, surface treatment, temperature and contact pressure are discussed. Contact between two Ge crystals is also considered.

242

621.314.63

The Characteristics and Applications of Metal Rectifiers.—P. A. Goodyer. (*Trans. S. Afr. Inst. elect. Engrs*, July 1949, Vol. 40, Part 7, pp. 147–172.) The characteristics of Se rectifiers are briefly described, and applications are considered fully. The relative merits of Se and Cu_2O rectifiers are discussed.

243

621.314.67

Thoriated Tungsten Filaments in Rectifiers.—Z. J. Atlee. (*Elect. Engng, N.Y.*, Oct. 1949, Vol. 68, No. 10, p. 863.) Summary only. A thoriated tungsten filament can have an emission efficiency of 30–40 mA/W at an operating temperature of 1 800–2 200 K with a life of 2 000 hr; it is thus superior to a pure tungsten filament. An account is given of the development of a very small 25-W rectifier for use in an X-ray application requiring a maximum of 110 kV and 250 mA from a full-wave bridge rectifier with 4 rectifier valves. Loss of emission by positive-ion bombardment is prevented by a new method of applying adequate getter. A new seasoning process prevents the deposit of thorium on the anode and glass.

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621.316.722.1

Operation of Voltage-Stabilizing Elements with Current-Stabilized Supplies.—J. J. Gilvarry & D. F. Rutland. (*Rev. sci. Instrum.*, Sept. 1949, Vol. 20, No. 9, pp. 633–637.) The performance parameters are determined. The range-regulation factor Q is defined and suggested as a figure of merit. Advantages of a current-stabilized power supply for voltage stabilization under varying load conditions are emphasized. Stability criteria for the case when the stabilizing element has a negative-resistance characteristic are given, and also a graphical analysis for the case of nonlinear elements. The special case of voltage-reference tubes is discussed, with experimental results.

245

621.396.68

Variable Frequency Power Supply.—E. B. Steinberg. (*Elect. Mfg, N.Y.*, May 1949, Vol. 43, No. 5, pp. 100–102.) Design details of a single-phase parallel inverter, using hydrogen thyratrons Type 5C22, which has an output of several kilowatts. The output frequency can be adjusted between 60 and 5 000 c/s. Performance data are discussed.

246

621.316.7.004.5 : 621.38

Maintenance Manual of Electronic Control. [Book Review]—R. E. Miller (Ed.). Publishers: McGraw-Hill, New York, 1949, 304 pp., $4.50. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 235–236.) Based on a series of articles published in *Electrical Construction and Maintenance* on the installation and service of electronic equipment. "The book should be a valuable aid to any engineer, maintenance man or technician who is involved in any way with timing relays, time-delay relays, photoelectric relays, electronic motor control, welding control, furnace temperature control and mercury-arc rectifiers."

TELEVISION AND PHOTOTELEGRAPHY

247

621.397.331.2

The Image Isocon — An Experimental Television Pickup Tube based on the Scattering of Low Velocity Electrons.—P. K. Weimer. (*RCA Rev.*, Sept. 1949, Vol. 10, No. 3, pp. 366–386.) A description of a new method of generating the video signal by introducing additional helical motion into the primary beam and collecting the electrons not reflected specularly to give an output signal of polarity opposite to that of the maximum current in the light. This results in (*a*) an improved signal/noise ratio in the darker parts of the picture, (*b*) freedom from spurious signals caused by multiplier dynode spots, an advantage which offsets the slightly superior resolution of the image orthicon. Closer tolerances in component design are required. The task of the operating crew is more exacting and the time lag is more objectionable than in the case of the image orthicon. Applications of the techniques used in the isocon are suggested for colour television and for image-storage tubes.

248

621.397.5

How RCA's Color TV Works.—E. W. Engstrom. (*FM-TV*, Oct. 1949, Vol. 9, No. 10, pp. 11–13, 30.) The principles of the system were discussed in 3297 of 1947 and 572 of 1948. Improvements have been added which make possible the transmission of a high-definition colour picture in a 6-Mc/s channel. Present receivers need no modification to receive these colour transmissions in monochrome. A block diagram of the broadcasting station is included and discussed. See also *Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 122–189.

249

621.397.5 : 535.88 : 791.45

Theater Television Today.—J. E. McCoy & H. P. Warner. (*J. Soc. Mot. Pict. Engrs*, Oct. 1949, Vol. 53, No. 4, pp. 321–350.)

250

621.397.5 : 621.396.81

WDTV Field-Strength Report.—T. T. Goldsmith, Jr. (*FM-TV*, Sept. 1949, Vol. 9, No. 9, pp. 15–18, 30.) Field-strength measurements for a hilly path radiating from Pittsburgh are tabulated, shown graphically, and compared with those expected from the results given in the Ad Hoc Committee's report (3524 of 1946).

251

621.397.5(437)

Television in Czechoslovakia.—J. Havelka. (*Tesla tech. Rep., Prague*, March 1949, pp. 2–3.) Post-war work has hitherto only been possible in laboratories. Only locally-produced equipment is used; this includes a kinescope with a flat square front. 625 lines, 25 frames/sec and negative modulation are used; picture and sound transmission are separate.

252

621.397.62 + 621.396.621

Mass Production of Radio and Television Receivers.—Alixant. (*See* 214.)

253

621.397.62

New Television Receiver without Transformers. Design with Interchangeable Units.—R. Aschen. (*T.S.F. pour Tous*, Oct. 1949, Vol. 25, No. 252, pp. 329–330.) Continuation of 2658 of 1949. Full details are given of the sound receiver, which uses two Type-UF41 pentodes as h.f. amplifier and grid detector respectively, and a Type-UL41 pentode for the output stage. Sensitivity is of the order of 200–300 μV for 50 mW in the loudspeaker.

254

621.397.62 : 621.385

Radio-Frequency Performance of some Receiving Valves in Television Circuits.—R. M. Cohen. (*Radiotronics*, July/Aug. 1949, No. 138, pp. 58–63.) Reprint. See 2971 of 1948.

621.397.645 255

Stagger-Tuned Television Amplifiers.—A. Easton. (*Radio tech. Dig., Edn franç.*, June 1949, Vol. 3, No. 3, pp. 147–152.) French version of 1344 of 1949.

621.397.7 256

WBAL-TV Channel 11, Baltimore.—W. C. Bareham. (*Broadcast News*, Sept. 1949, No. 56, pp. 72–80.) General description of facilities available, with photographs of studios, control room, etc., and map showing coverage.

621.397.8 257

TV Reception below Line of Sight.—R. B. McGregor. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 72–76.) Signals were received at a point 92 miles from the transmitter and 2 000 ft below the line of sight, with the aid of an 18-element aerial array, a cascode pre-amplifier (3065 of 1948) at the aerial and special i.f. amplifier stages and sweep circuits. Signals were received whenever the transmitter was operating. About 50% of the time, the picture was satisfactory. Sometimes it remained consistently good all the evening. Sometimes it was good by day and poor by night, more often vice versa. Some frequency-selective reception was noted. No correlation between reception and weather could be found.

621.397.82 258

The Influence of U.H.F. Allocations on Receiver Design.—J. D. Reid. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1179–1181.) Added protection of receivers from local oscillator radiation can be obtained by alternate channel assignments and receiver i.f. standardization. Additional protection from image interference can be given by having the picture-signal station separated by $\sqrt{2}$ times the distance between stations on adjacent channels, and co-channel stations separated by double this distance. 41.25 Mc/s is suggested as the optimum i.f. for joint v.h.f./u.h.f. usage.

TRANSMISSION

621.396.61 : 621.396.662 259

A New Type of V.H.F. Tank Design.—Parker. (*See* 79.)

621.396.619.23 260

Rectifier Modulators with Frequency-Selective Terminations.—D. G. Tucker. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 422–428.) A simple method is given for determining the performance of such a modulator provided that the terminating impedance is resistive, zero or infinite at all significant frequencies. Cases likely to be useful or unavoidable in practice are worked out for the ring, Cowan (shunt-type) and series modulators. Optimum terminating resistances and minimum insertion losses are tabulated in terms of the rectifier ratio and geometric-mean resistance.

VALVES AND THERMIONICS

621.385 : 621.396.645 261

Operation of Output Valves in High-Power Public-Address Amplifiers.—Bezladnov. (*See* 75.)

621.385 : 621.397.62 262

Radio-Frequency Performance of some Receiving Valves in Television Circuits.—R. M. Cohen. (*Radiotronics*, July/Aug. 1949, No. 138, pp. 58–63.) Reprint. See 2071 of 1948.

621.385.001.4 : 621.3.015.3 263

Surge Testing of High Vacuum Tubes.—Dailey. (*See* 169.)

621.385.029.63/64 : 621.317.755 264

Traveling-Wave Oscilloscope.—J. R. Pierce. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 97–99.) The 1 000-V oscilloscope here described was developed specially for laboratory examination of short recurrent pulses. Response is almost flat from 0 to 500 Mc/s. Input impedance is 75Ω. A peak-to-peak signal of 0.37 V gives a pattern 10 trace-widths high, which is viewed through a microscope. Good vertical deflection sensitivity is obtained by means of a travelling-wave deflection system, without transit-time bandwidth limitations. The essentials of the oscilloscope are illustrated and discussed.

621.385.032.21 : 621.317.39 : 536.5 265

On a New Method of Measuring the Temperature of a Thermionic Cathode.—P. Gandin & R. Champeix. (*C. R. Acad. Sci., Paris*, 12th Sept. 1949, Vol. 229, No. 11, pp. 545–547.) By differentiation of the expression for the anode current I in terms of the saturation current, negative voltage of the anode, and cathode temperature T, the following formula is derived: $T = 11\,600\rho I$, where ρ is the differential resistance of the cathode/anode space in ohms, I is in amperes and T in absolute degrees. A simple measurement circuit is described; this makes use of a transformer whose secondary provides two equal voltages of opposite phase. These voltages should be less than 0.01 V. One voltage is applied to cathode and anode, a variable d.c. voltage and a galvanometer being included in the cathode lead; the other voltage feeds a standard resistor, one terminal of which is connected to the anode. A null indicator is connected in the lead common to the two branches. Errors in determining T should not exceed 2%. Results obtained by this method will be given in a later paper.

621.385.032.212 : 681.142 266

Polycathode Glow Tube for Counters and Calculators.—J. J. Lamb & J. A. Brustman. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 92–96.) A cold-cathode neon-filled discharge tube whose basket-shaped anode has 30 narrow slots. Cathode fingers, in staggered sets of 10, are located so as to line up with the anode segments, the gap being 0.020 in. The 10th finger of one set is separate from the remainder and is used for numerical carrying. The initial potential difference between anode and cathodes is such that a discharge takes place between only one of the cathode fingers and the anode when a potential exceeding the breakdown voltage is applied. The residual ionization around this finger favours the formation of the next discharge at the adjacent finger of another cathode ring when that ring is energized. The tube is capable of operation at rates up to 10^8 per sec and can be adapted to decade counting circuits with rates exceeding 16 000 per sec.

621.385.832 : 535.371.07 : 621.396.9 267

Radar Screens.—B. de la Pinsonie. (*Bull. Soc. franç. Élect.*, Oct. 1949, Vol. 9, No. 97, pp. 532–542.) The mechanism of fluorescence and phosphorescence is discussed, the characteristics of screens specially adapted for radar presentation are considered and suitable methods for measuring the various characteristics are described, with experimental results, both for ordinary radar screens and for the skiatron type.

MISCELLANEOUS

621.396 268

Reference Data for Radio Engineers. [Book Review]—Publishers: Federal Telephone and Radio Corporation, New York, 3rd edn 1949, 672 pp., $3.75. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, p. 242.) Material in the second edition (see 1301 of 1947) has been expanded and new material added, doubling the size. See also 236 above.

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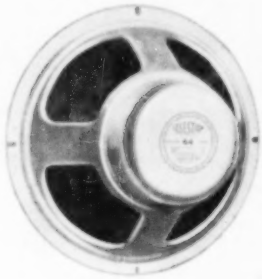

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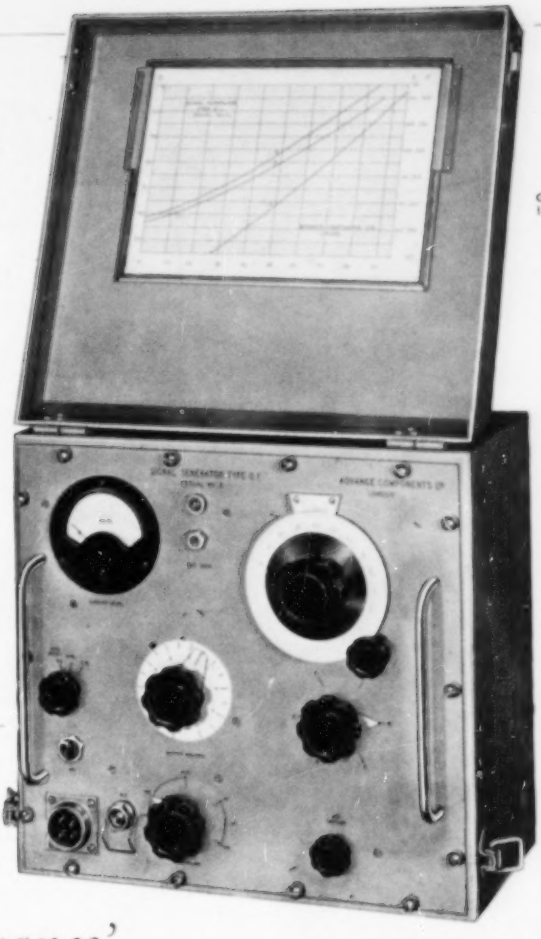

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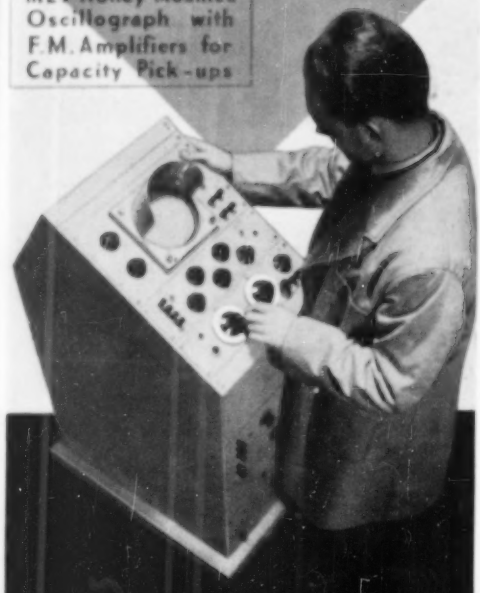

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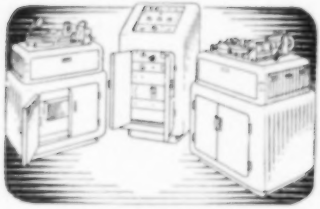

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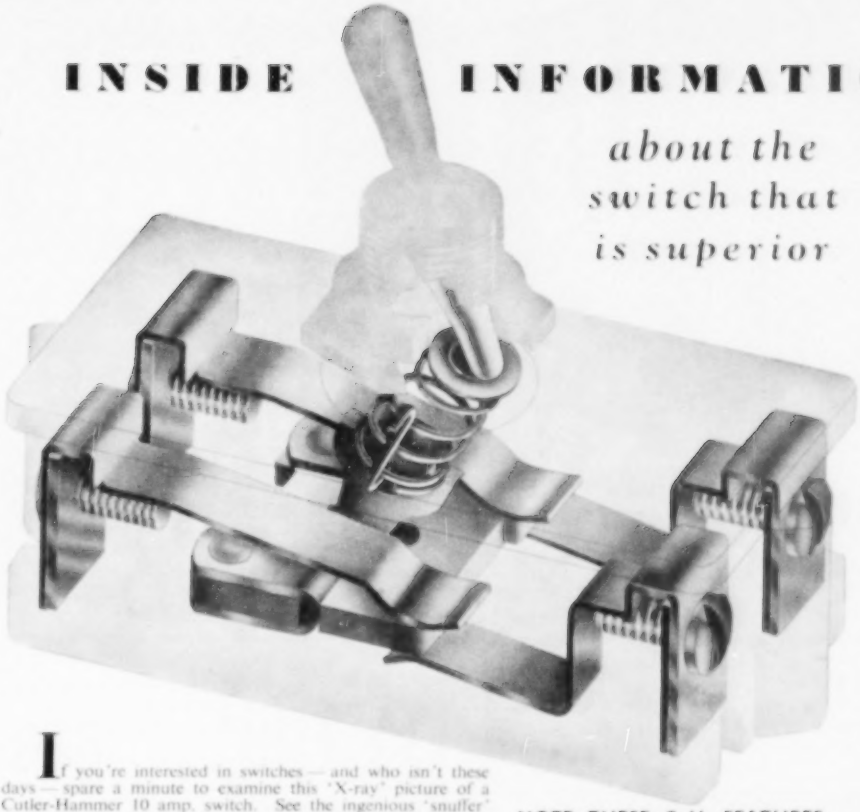

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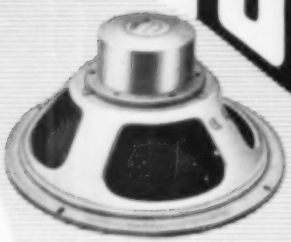

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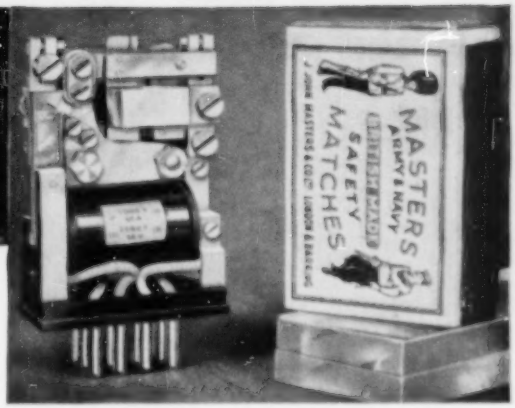

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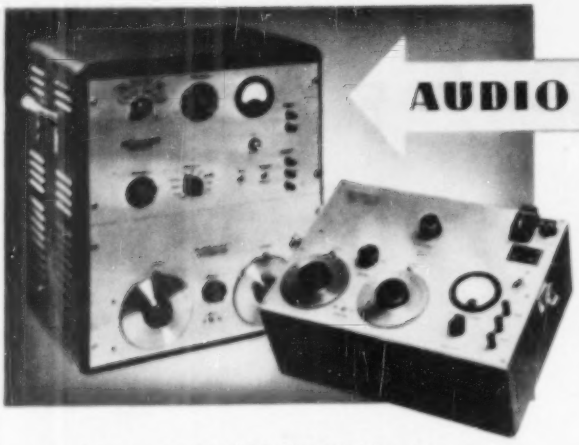

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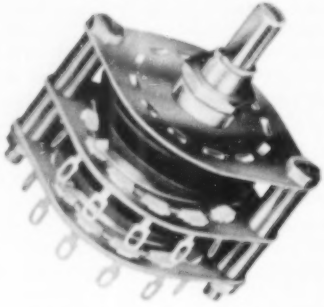

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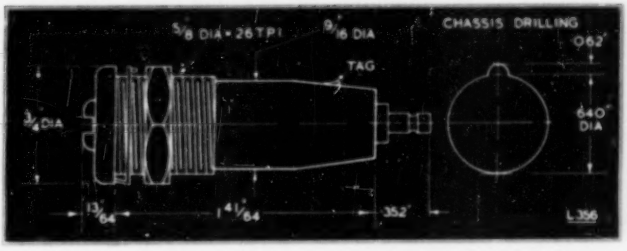

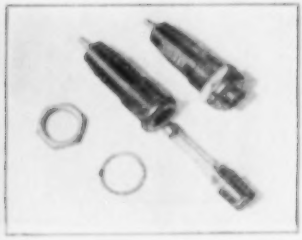

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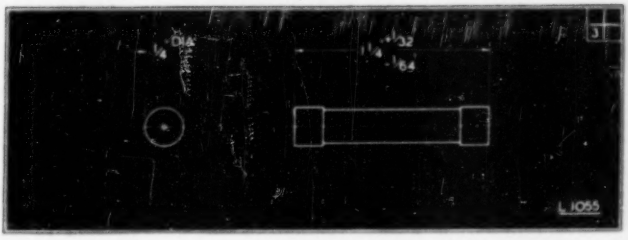

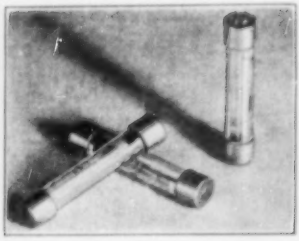

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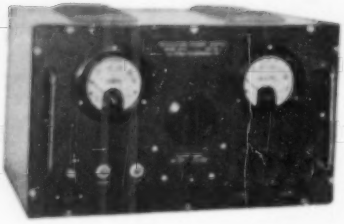

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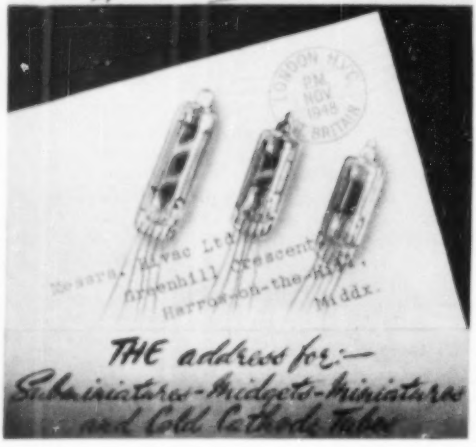

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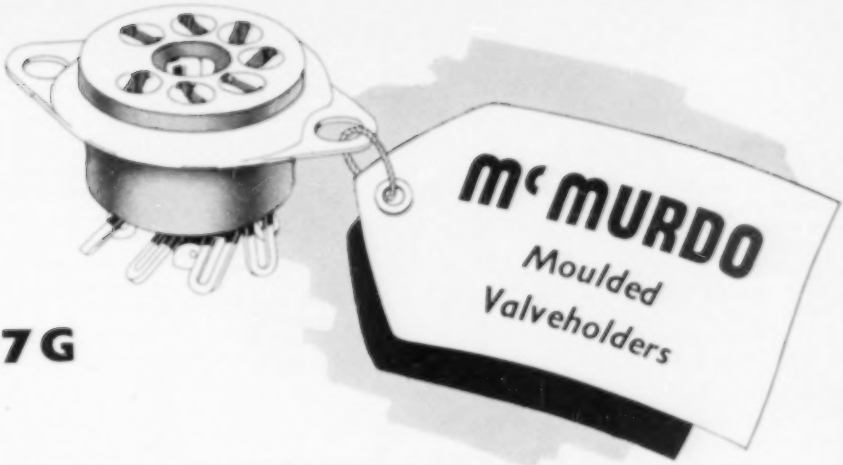

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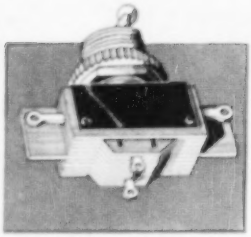

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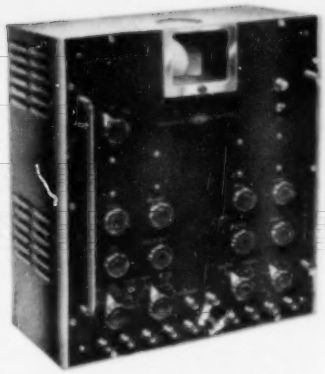

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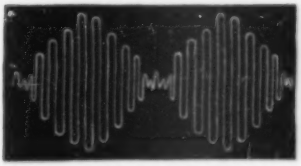

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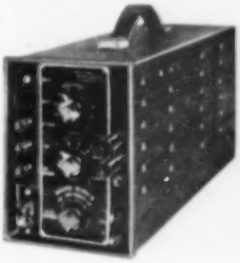

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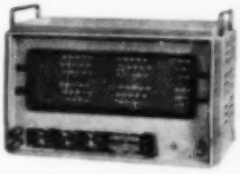

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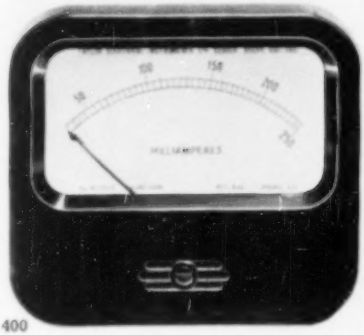

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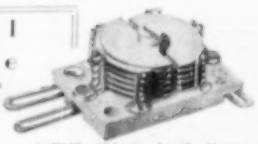

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·001	18000	5⅛	1⅛	CP57XO	13/6
·002	18000	5⅛	⅞	CP55HO	4/6
·0025	3000	2⅛	⅞	CP56QO	7/6
·005	6000	3	1⅛	CP56QO	7/6
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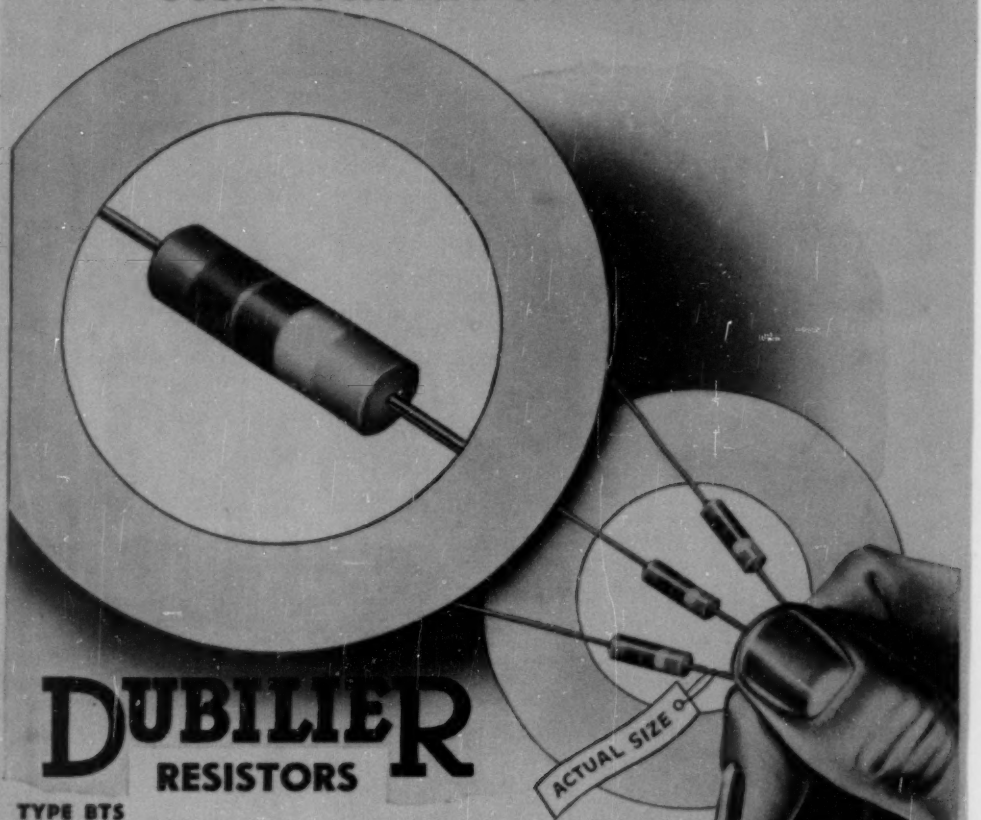

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